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- 10 effects to choose, including echo, reverb, vibrato and overdrive
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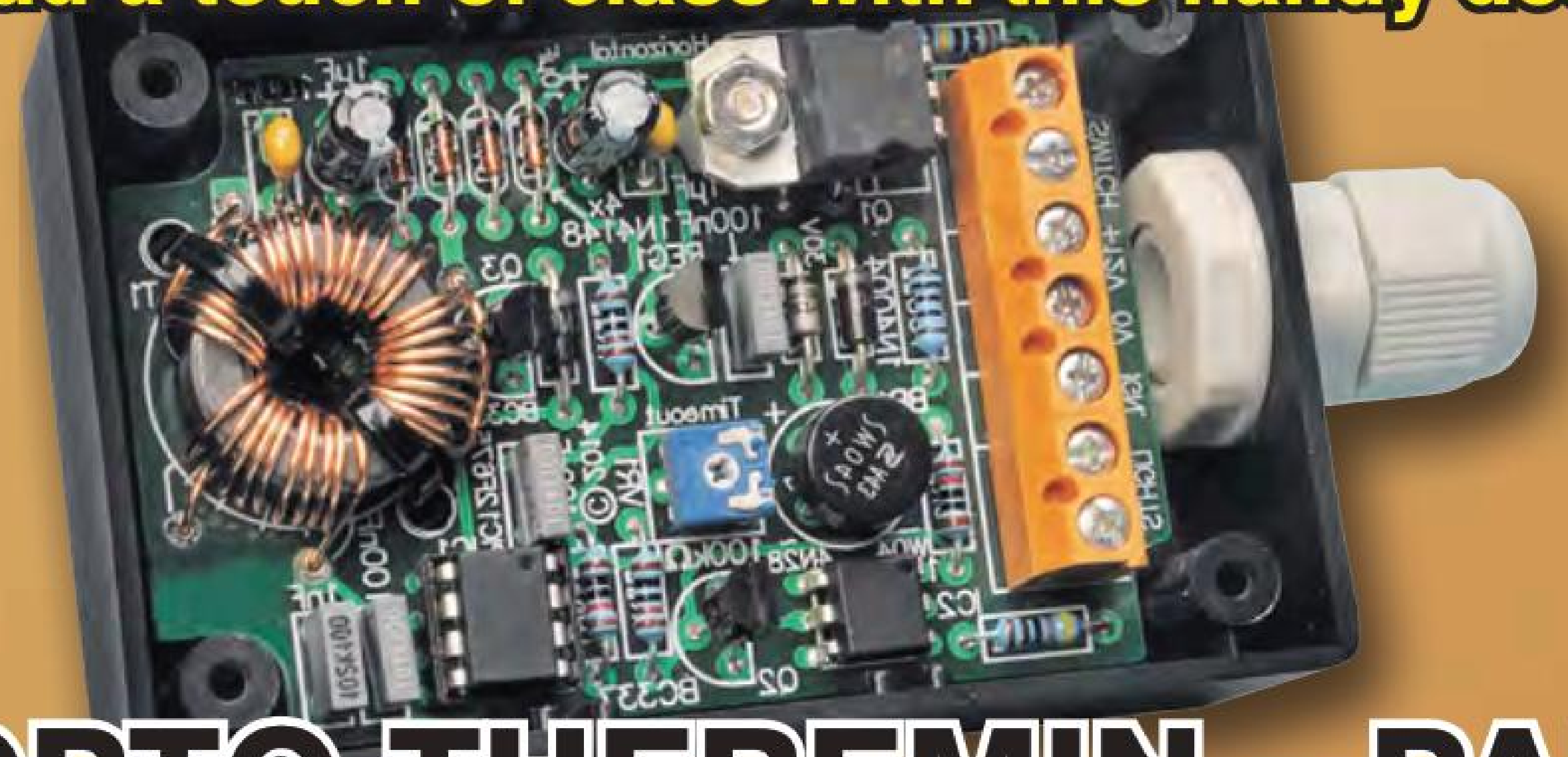

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Construction, testing and adjustment

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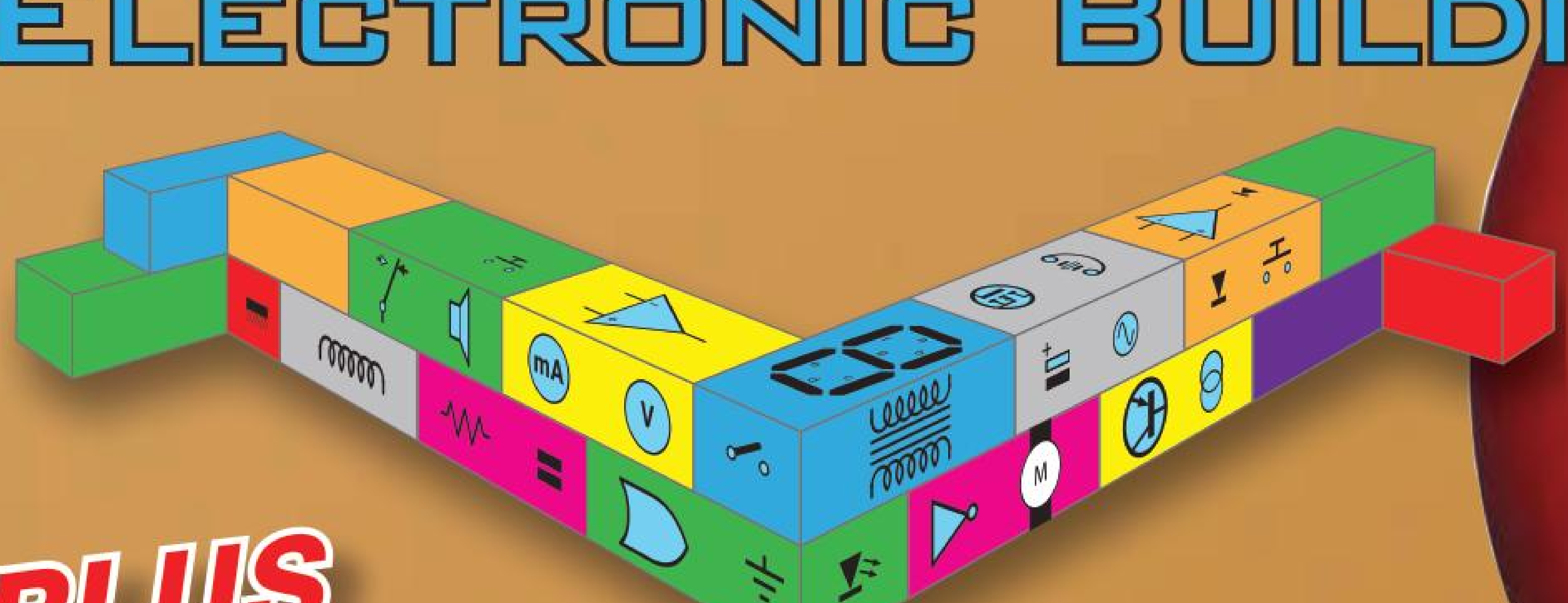

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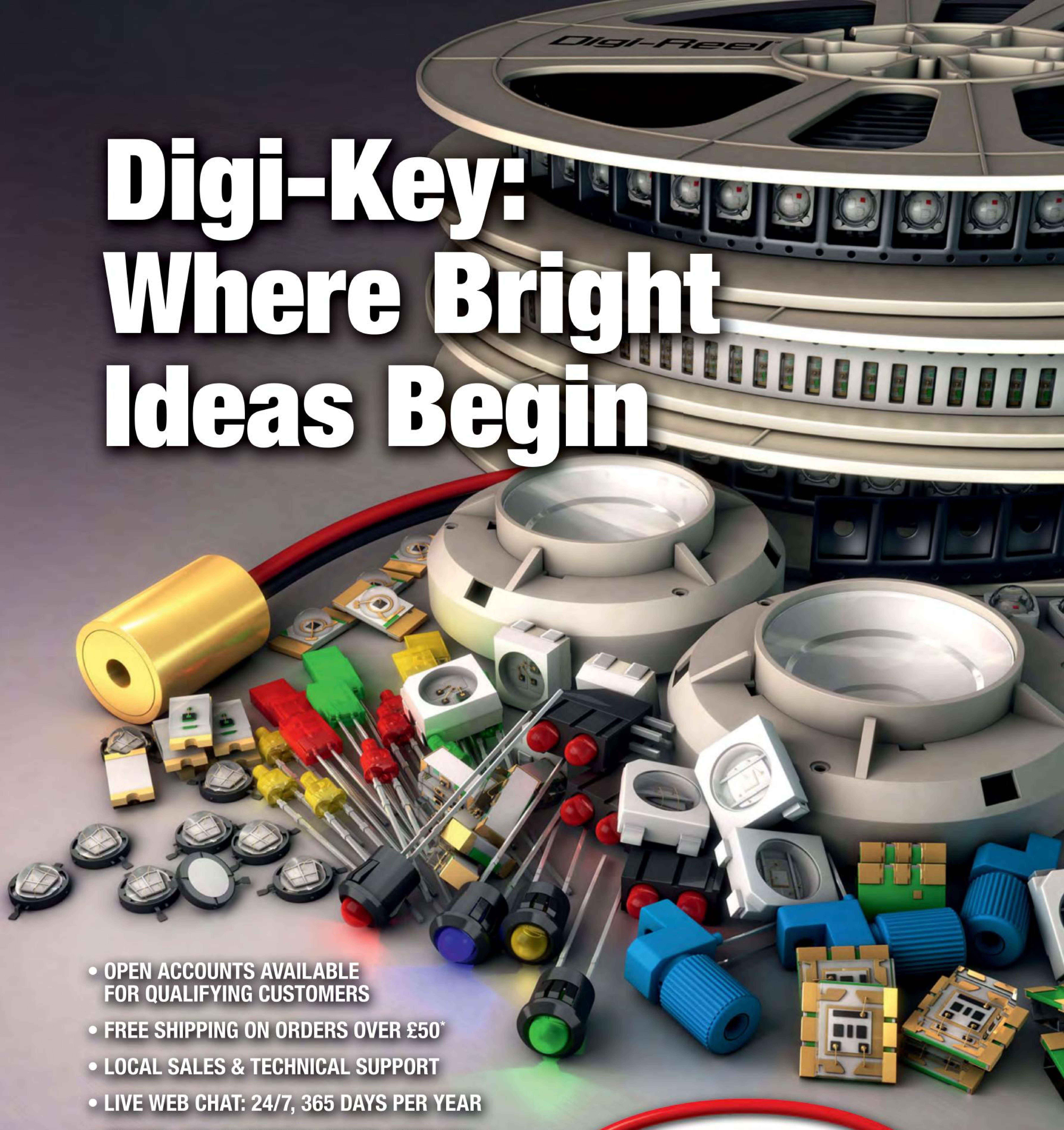

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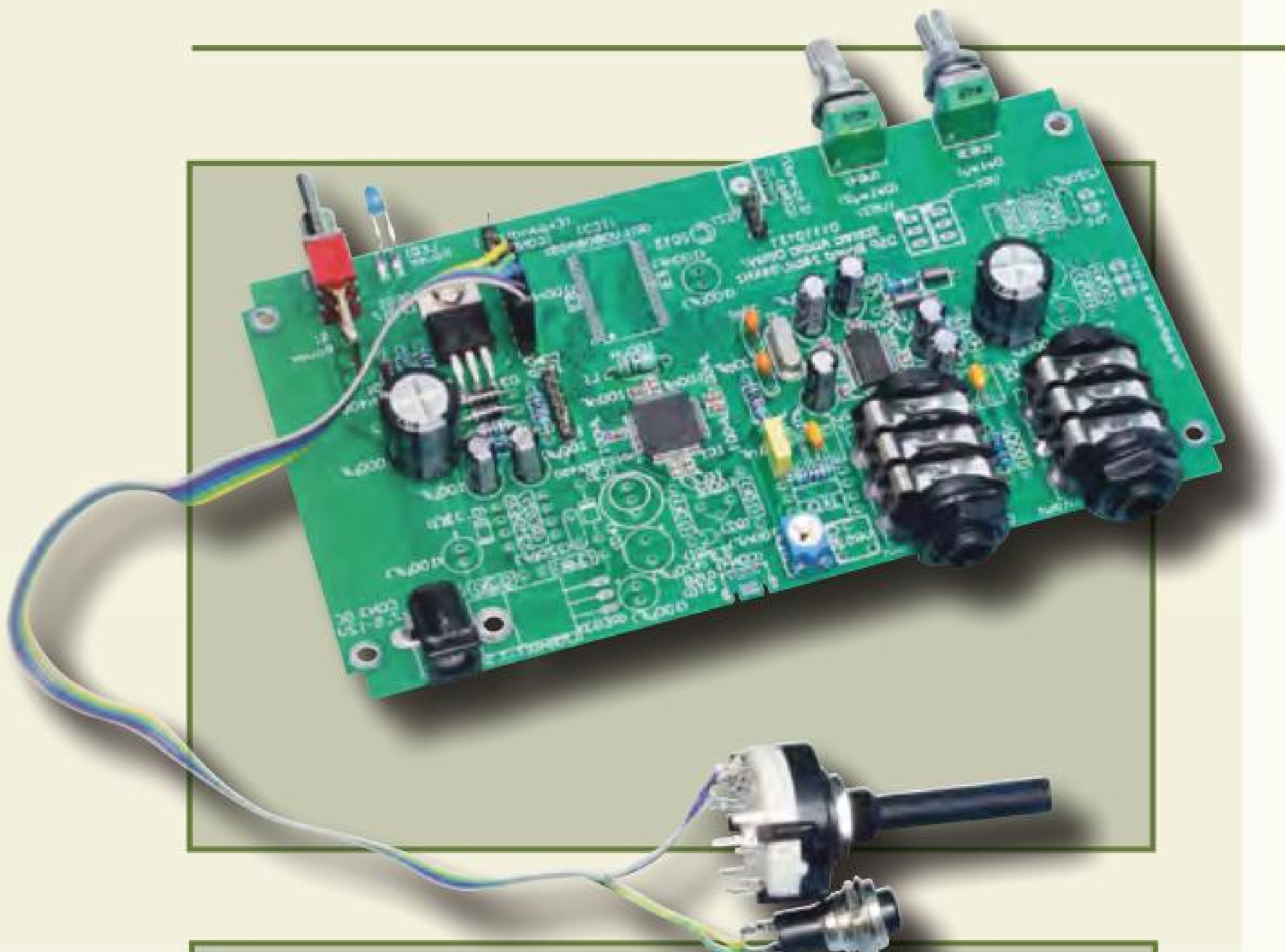

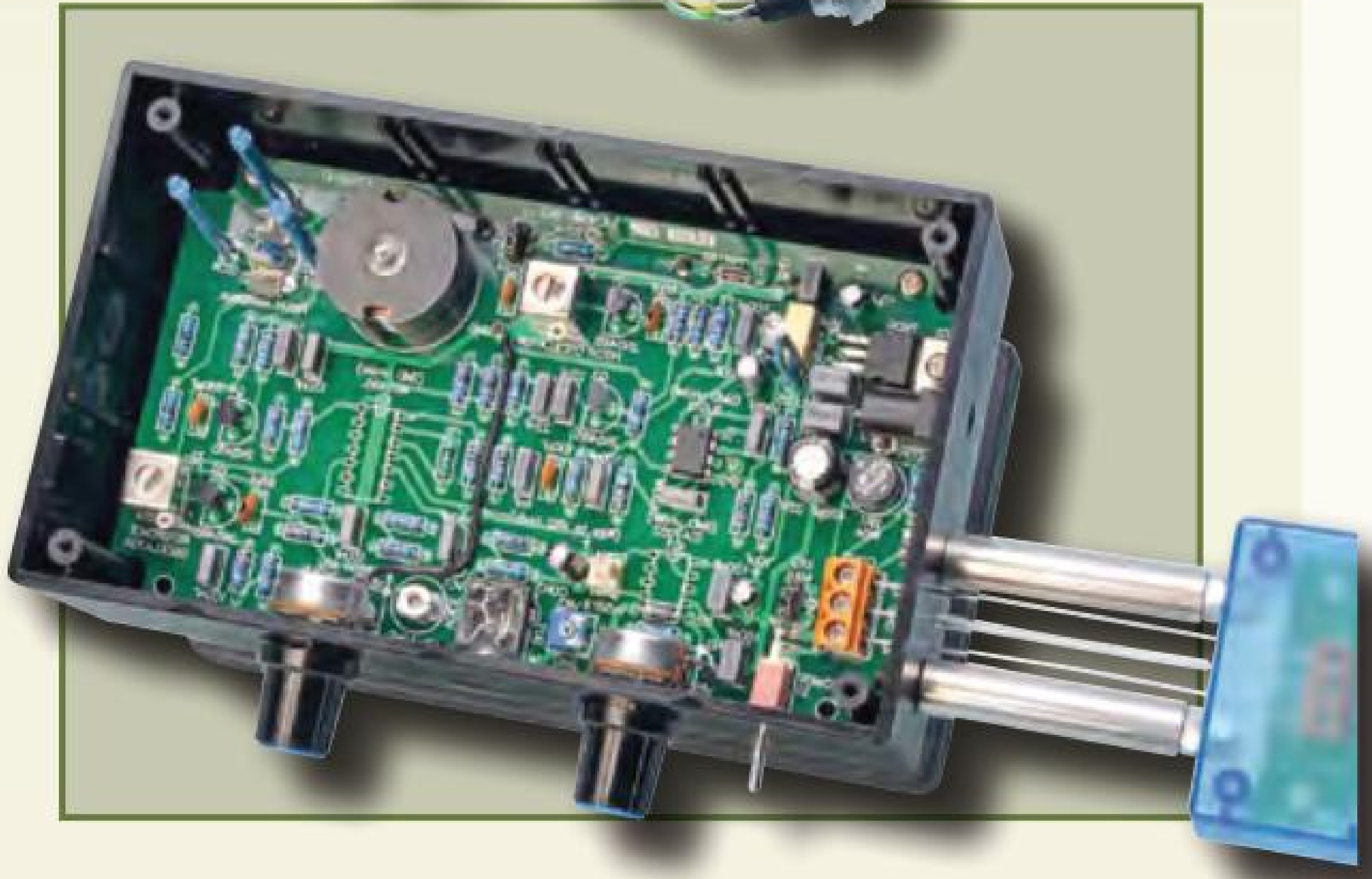

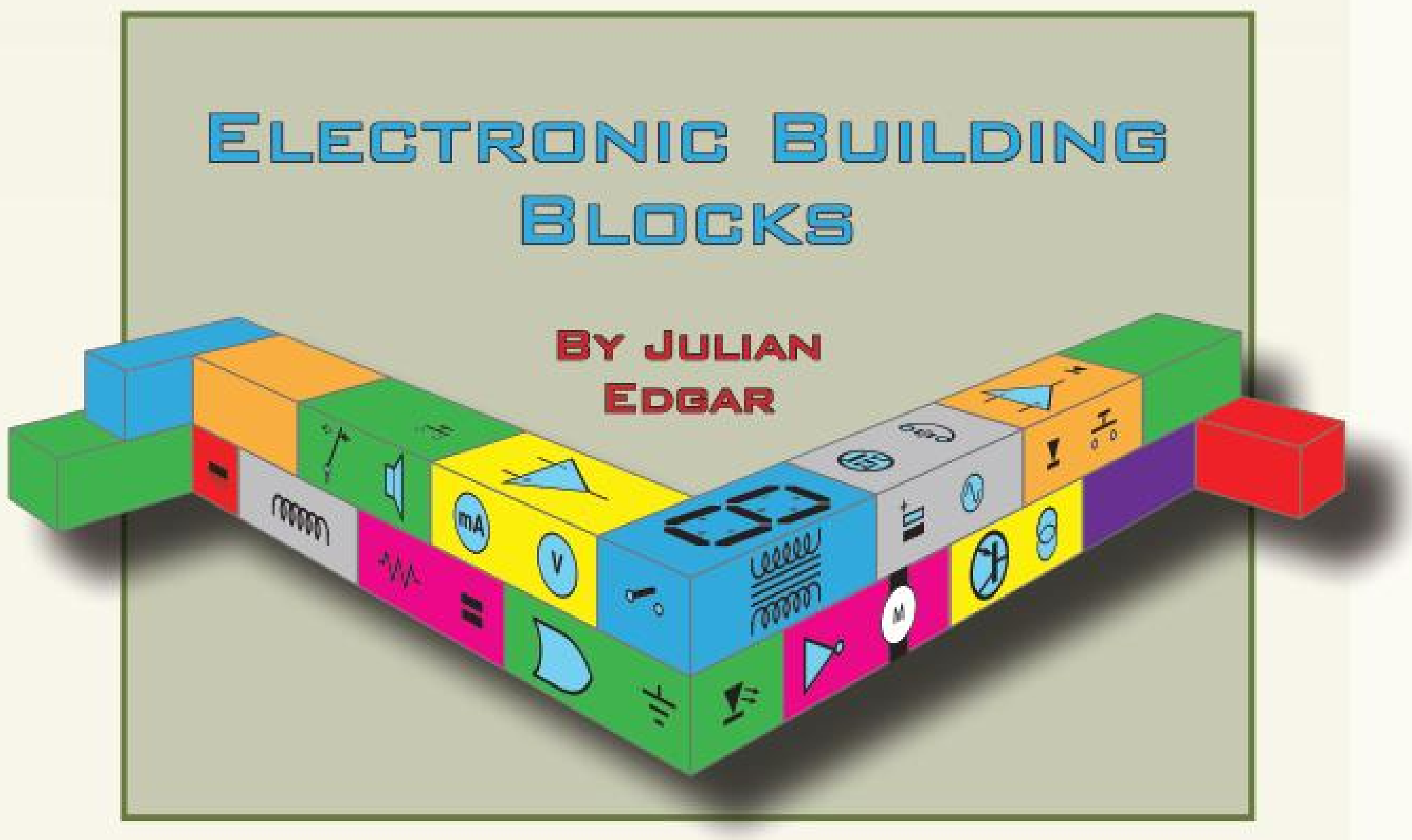

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Our November 2015 issue will be published on Thursday 1 October 2015, see page 72 for details.

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Assembled with ZIF socket Order Code: AS3149EZIF - £74.95

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This tutorial project board is all you need to take your first steps into Microchip PIC programming using a PIC16F882 (included). Later you can use it for more advanced programming. It programs all the devices a Microchip PICKIT2® can! You can use the free Microchip tools for the PICKit2™ and the MPLAB® IDE environment.
*Order Code: EDU10 - **£55.96***

ATMEL 89xxxx Programmer

Uses serial port and any standard terminal comms program. 4 LED's display the status. ZIF sockets not included. 16Vdc.
*Kit Order Code: 3123KT - **£28.95***
*Assembled Order Code: AS3123 - **£39.95***

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Go from complete beginner to burning a PIC and writing code in no time! Includes 49 page step-by-step PDF Tutorial Manual + Programming Hardware (with LED test section) + Windows Software (Program, Read, Verify & Erase) + a rewritable PIC16F84A. 4 detailed examples provided for you to learn from. PC parallel port. 12Vdc.
*Kit Order Code: 3081KT - **£16.95***
*Assembled Order Code: AS3081 - **£24.95***

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Low cost PIC programmer board supporting a wide range of Microchip® PIC™ microcontrollers. Serial port. Free Windows software.
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PIC Programmer & Experimenter Board with test buttons and LED indicators to carry out educational experiments such as the supplied programming examples. Includes a 16F627 Flash Microcontroller that can be reprogrammed up to 1000 times. Software to compile and program your source code is included. Supply: 12-15Vdc.
*Kit Order Code: K8048 - **£23.94***
*Assembled Order Code: VM111 - **£39.12***

Controllers & Loggers

Here are just a few of the controller and data acquisition and control units we have. See website for full details. 12Vdc PSU for all units: Order Code 660.446UK £11.52

USB Experiment Interface Board

5 digital input channels and 8 digital output channels plus two analogue inputs and two analogue outputs with 8 bit resolution.
*Kit Order Code: K8055N - **£25.19***
*Assembled Order Code: VM110N - **£40.20***

2-Channel High Current UHF RC Set

State-of-the-art high security. 2 channel. Momentary or latching relay output rated to switch up to 240Vac @ 10 Amps. Range up to 40m. Up to 15 Tx's can be learnt by one Rx (kit includes one Tx but more available separately). 3 indicator LEDs. Rx: PCB 88x60mm, supply 9-15Vdc.
*Kit Order Code: 8157KT - **£49.95***
*Assembled Order Code: AS8157 - **£54.95***

Computer Temperature Data Logger

Serial port 4-channel temperature logger. °C or °F. Continuously logs up to 4 separate sensors located 200m+ from board. Wide range of free software applications for storing/using data. PCB just 45x45mm. Powered by PC. Includes one DS1820 sensor.
*Kit Order Code: 3145KT - **£19.95***
*Assembled Order Code: AS3145 - **£26.95***
*Additional DS1820 Sensors - **£4.95 each***

Remote Control Via GSM Mobile Phone

Place next to a mobile phone (not included). Allows toggle or autotimer control of 3A mains rated output relay from any location

Most items are available in kit form (KT suffix) or pre-assembled and ready for use (AS prefix).

4-Ch DTMF Telephone Relay Switcher

Call your phone number using a DTMF phone from anywhere in the world and remotely turn on/off any of the 4 relays as desired. User settable Security Password, Anti-Tamper, **Rings** to Answer, Auto Hang-up and Lockout. Includes plastic case. 130 x 110 x 30mm. Power: 12Vdc.
*Kit Order Code: 3140KT - **£79.95***
*Assembled Order Code: AS3140 - **£94.95***

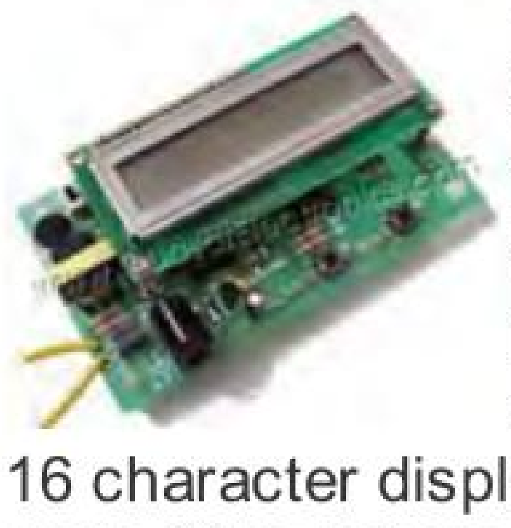
8-Ch Serial Port Isolated I/O Relay Module

Computer controlled 8 channel relay board. 5A mains rated relay outputs and 4 opto-isolated digital inputs (for monitoring switch states, etc). Useful in a variety of control and sensing applications. Programmed via serial port (use our new Windows interface, terminal emulator or batch files). Serial cable can be up to 35m long. Includes plastic case 130x100x30mm. Power: 12Vdc/500mA.
*Kit Order Code: 3108KT - **£74.95***
*Assembled Order Code: AS3108 - **£89.95***

Infrared RC 12–Channel Relay Board

Control 12 onboard relays with included infrared remote control unit. Toggle or momentary. 15m+ range. 112 x 122mm. Supply: 12Vdc/0.5A
*Kit Order Code: 3142KT - **£64.95***
*Assembled Order Code: AS3142 - **£74.95***

Audio DTMF Decoder and Display



Detect DTMF tones from tape recorders, receivers, two-way radios, etc using the built-in mic *or* direct from the phone line. Characters are displayed on a 16 character display as they are received and up to 32 numbers can be displayed by scrolling the display. All data written to the LCD is also sent to a serial output for connection to a computer. Supply: 9-12V DC (Order Code PSU375). Main PCB: 55x95mm.
*Kit Order Code: 3153KT - **£37.95***
*Assembled Order Code: AS3153 - **£49.95***

3x5Amp RGB LED Controller with RS232

3 independent high power channels. Preprogrammed or user-editable light sequences. Standalone option and 2-wire serial interface for microcontroller or PC communication with simple command set. Suitable for common anode RGB LED strips, LEDs and incandescent bulbs. 56 x 39 x 20mm. 12A total max. Supply: 12Vdc.
Kit Order Code: 8191KT - £29.95
Assembled Order Code: AS8191 - £39.95

Hot New Products!

Here are a few of the most recent products added to our range. See website or join our email Newsletter for all the latest news.

4-Channel Serial Port Temperature Monitor & Controller Relay Board

4 channel computer serial port temperature monitor and relay controller. Four inputs for Dallas DS18S20 or DS18B20 digital thermometer sensors (£3.95 each). Four 5A rated relay outputs are independent of sensor channels allowing flexibility to setup the linkage in any way you choose. Simple text string commands for reading temperature and relay control via RS232 using a comms program like Windows HyperTerminal or our free Windows application.
*Kit Order Code: 3190KT - **£84.95***
*Assembled Order Code: AS3190 - **£99.95***

40 Second Message Recorder

Feature packed non-volatile 40 second multi-message sound recorder module using a high quality Winbond sound recorder IC. Standalone operation using just six onboard buttons or use onboard SPI interface. Record using built-in microphone or external line in. 8-24Vdc powered. Change a resistor for different recording duration/sound quality. Sampling frequency 4-12 kHz. (120 second version also available)
*Kit Order Code: 3188KT - **£29.95***
*Assembled Order Code: AS3188 - **£37.95***

Bipolar Stepper Motor Chopper Driver

Get better performance from your stepper motors with this dual full bridge motor driver based on SGS Thompson chips L297 & L298. Motor current for each phase set using on-board potentiometer. Rated to handle motor winding currents up to 2 Amps per phase. Operates on 9-36Vdc supply voltage. Provides all basic motor controls including full or half stepping of bipolar steppers and direction control. Allows multiple driver synchronisation. Perfect for desktop CNC applications.
*Kit Order Code: 3187KT - **£39.95***
*Assembled Order Code: AS3187 - **£49.95***

Video Signal Cleaner

Digitally cleans the video signal and removes unwanted distortion in video signal. In addition it stabilises picture quality and luminance fluctuations. You will also benefit from improved picture quality on LCD monitors or projectors.
*Kit Order Code: K8036 - **£24.70***
*Assembled Order Code: VM106 - **£36.53***

Motor Speed Controllers

Here are just a few of our controller and driver modules for AC, DC, Unipolar/Bipolar stepper motors and servo motors. See website for full details.

DC Motor Speed Controller (100V/7.5A)

Control the speed of almost any common DC motor rated up to 100V/7.5A. Pulse width modulation output for maximum motor torque at all speeds. Supply: 5-15Vdc. Box supplied. Dimensions (mm): 60Wx100Lx60H.
*Kit Order Code: 3067KT - **£19.95***
*Assembled Order Code: AS3067 - **£27.95***

Bidirectional DC Motor Speed Controller

Control the speed of most common DC motors (rated up to 32Vdc/10A) in both the forward and reverse direction. The range of control is from fully OFF to fully ON in both directions. The direction and speed are controlled using a single potentiometer. Screw terminal block for connections.
*Kit Order Code: 3166v2KT - **£23.95***
*Assembled Order Code: AS3166v2 - **£33.95***

Computer Controlled / Standalone Unipolar Stepper Motor Driver

Drives any 5-35Vdc 5, 6 or 8-lead unipolar stepper motor rated up to 6 Amps. Provides speed and direction control. Operates in stand-alone or PC-controlled mode for CNC use. Connect up to six 3179 driver boards to a single parallel port. Board supply: 9Vdc. PCB: 80x50mm.
*Kit Order Code: 3179KT - **£17.95***
*Assembled Order Code: AS3179 - **£24.95***

Computer Controlled Bi-Polar Stepper Motor Driver

Drive any 5-50Vdc, 5 Amp bi-polar stepper motor using externally supplied 5V levels for STEP and DIRECTION control. Opto-isolated inputs make it ideal for CNC applications using a PC running suitable software. Board supply: 8-30Vdc. PCB: 75x85mm.
*Kit Order Code: 3158KT - **£24.95***
*Assembled Order Code: AS3158 - **£34.95***

AC Motor Speed Controller (600W)

Reliable and simple to install project that allows you to adjust the speed of an electric drill or 230V AC single phase induction motor rated up to 600 Watts. Simply turn the potentiometer to adjust the motors RPM. PCB: 48x65mm. Not suitable for use with brushless AC motors.
*Kit Order Code: 1074KT - **£15.95***
*Assembled Order Code: AS1074 - **£23.95***

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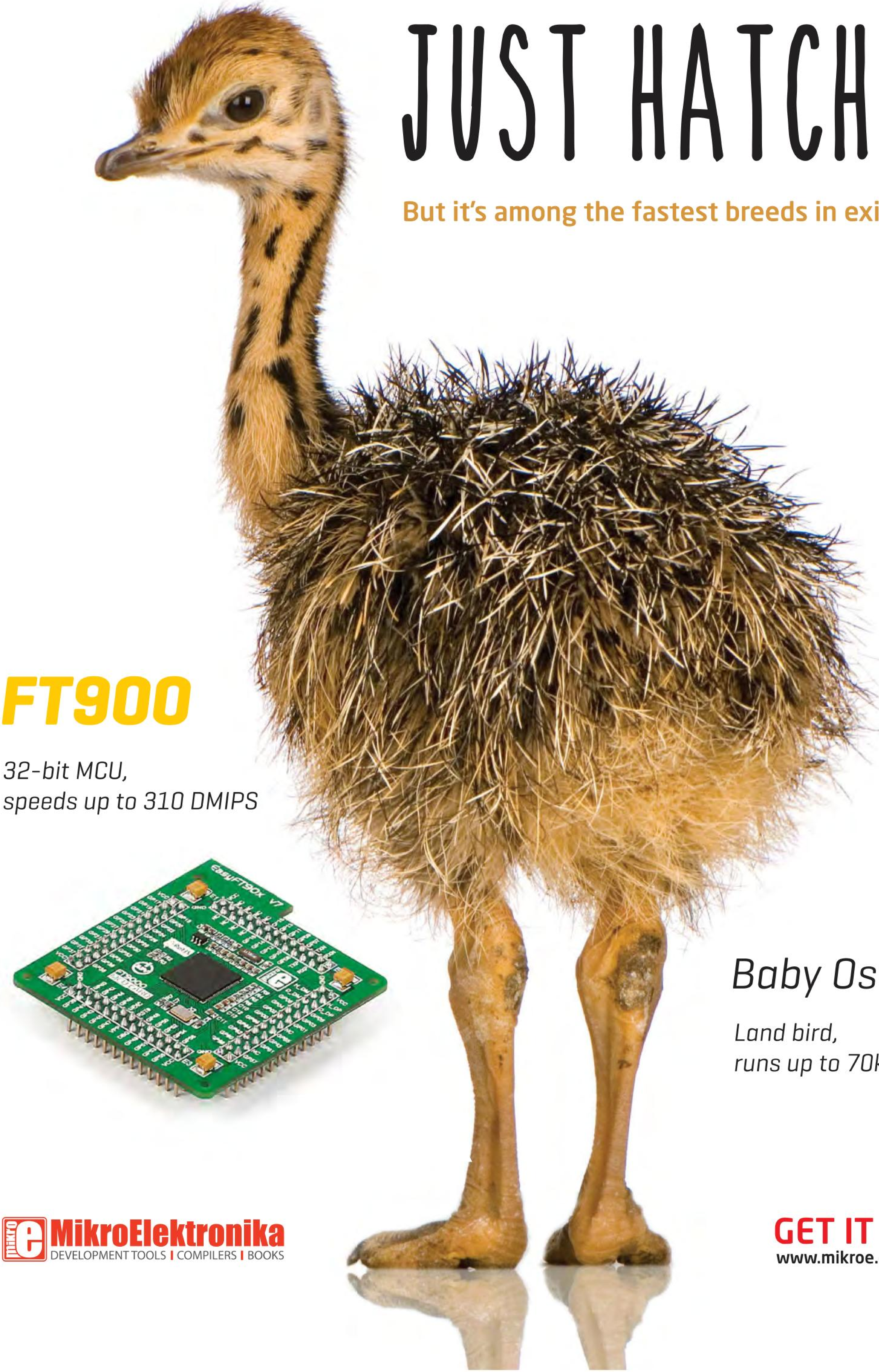

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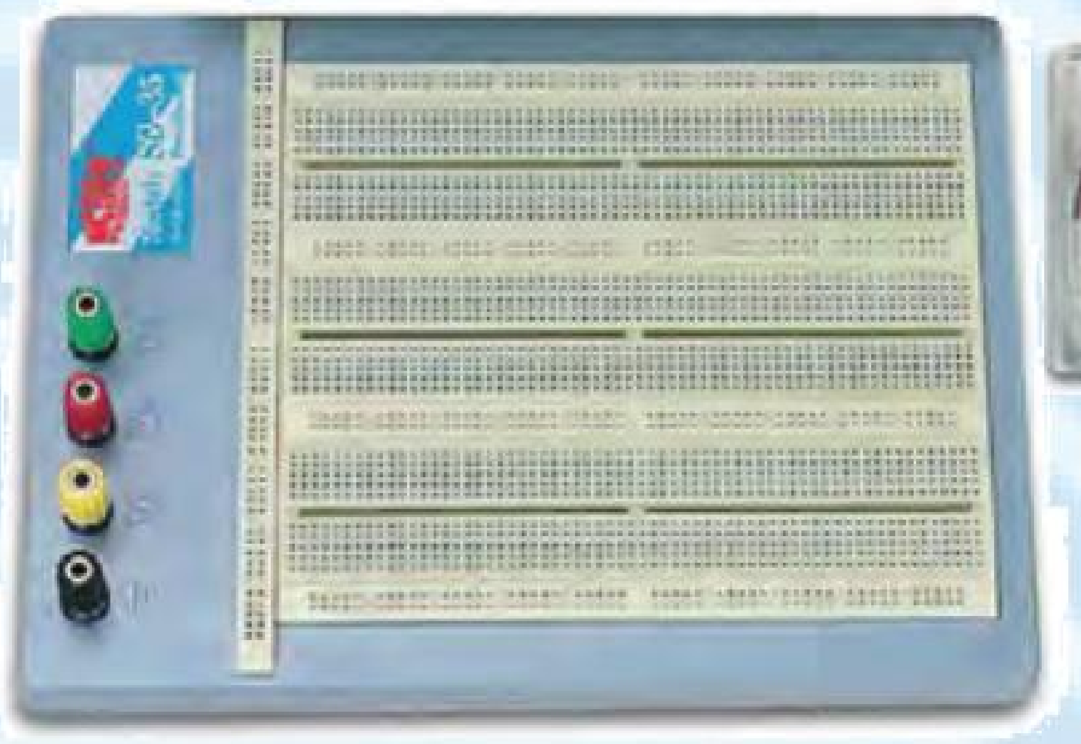

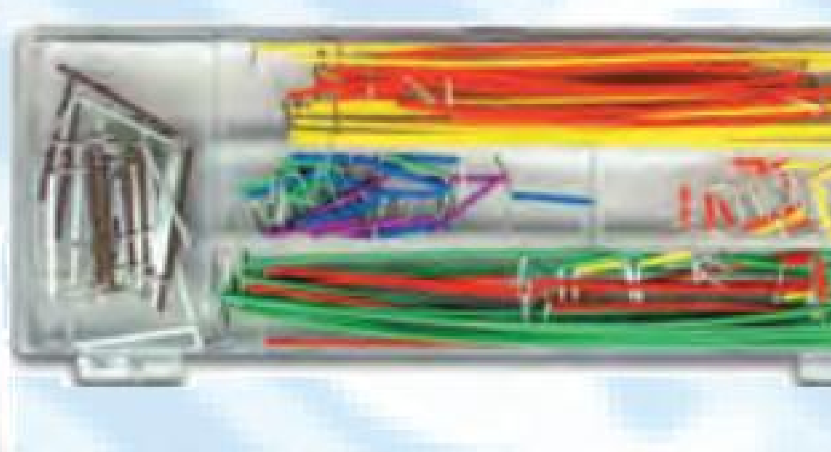

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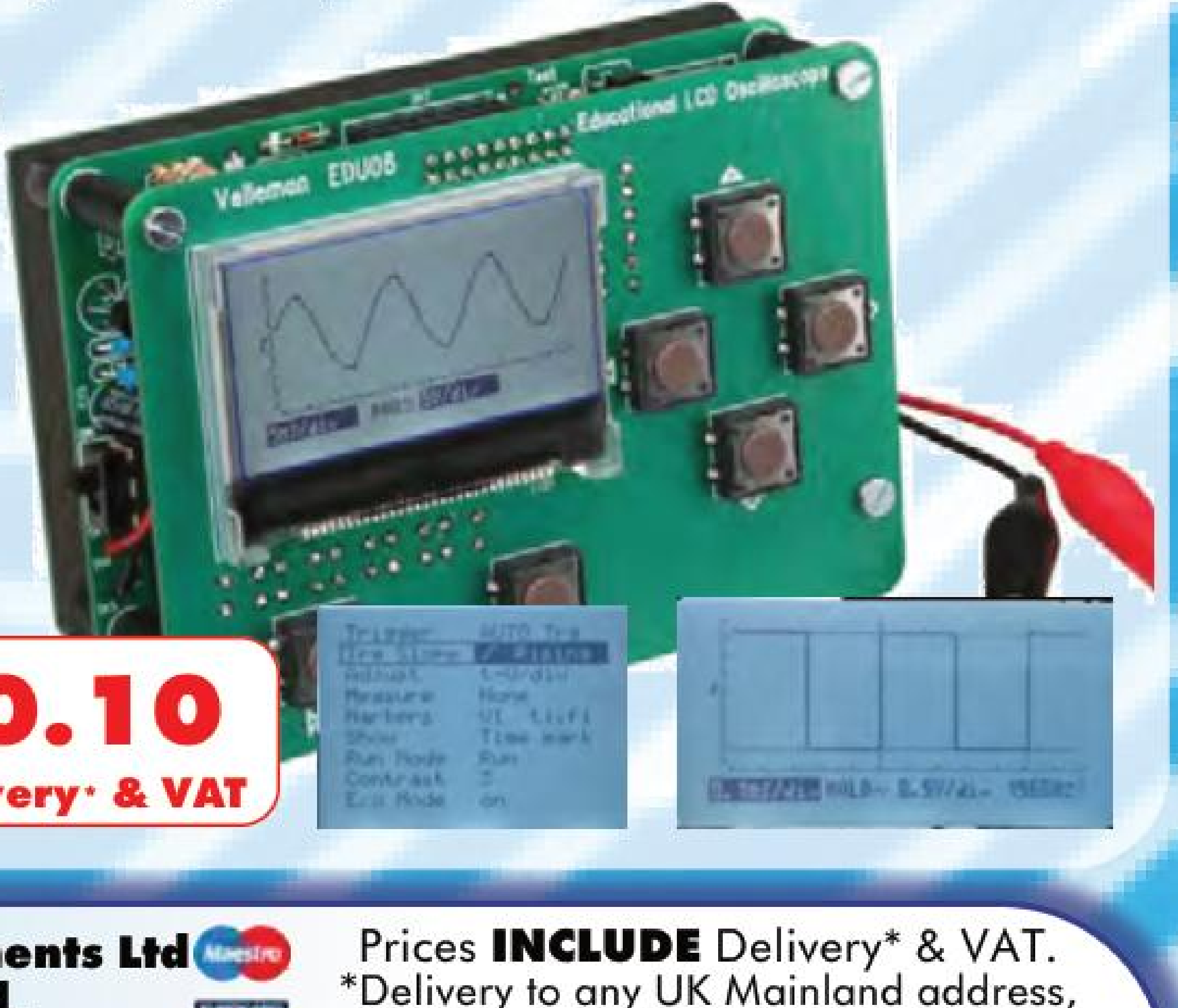

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EPE EVERYDAY PRACTICAL ELECTRONICS

Putting our 'regulars' in the spotlight!

I often focus on projects in my editorials, but that does rather sell short our excellent columnists. So this month, I have decided to focus on our regular features. We have some fascinating articles for you in this issue. Mark Nelson's *Techno Talk* provides some illuminating insights into the latest light-based technology – from solar power to novel LED fabrication technology.

Net Work is the usual must-read compendium of news, technology, tips and advice. I do wonder how Alan finds time to cram all his activities into what must be a very busy schedule that includes looking after our website, helping *Chat Zone* forum members and writing *Net Work*.

Continuing his *Interface* series using the popular Raspberry Pi, Robert Penfold looks at the powerful technique of pulse-width modulation (PWM). From electric motors to LEDs, this elegant and simple technique is a great way to control the analogue world with digital techniques.

Ian Bell's *Circuit Surgery* is a wonderful combination of in-depth analysis coupled with practical down-to-earth design, and this month focuses on current mirrors; helping a reader understand why real-word circuits can depart from their idealised versions. Naturally, LTSpice is on hand to take the drudgery out of endless calculations and error-prone manual modelling. If you have yet to take the plunge with LTSpice then you really should promise yourself to make its acquaintance, it is an absorbing and invaluable tool.

In *Audio Out*, Jake Rothman has completed his four-part series on RIAA filters – even if audio is not your passion, Jakes's columns are well worth reading for their tips and insight into high-quality analogue design.

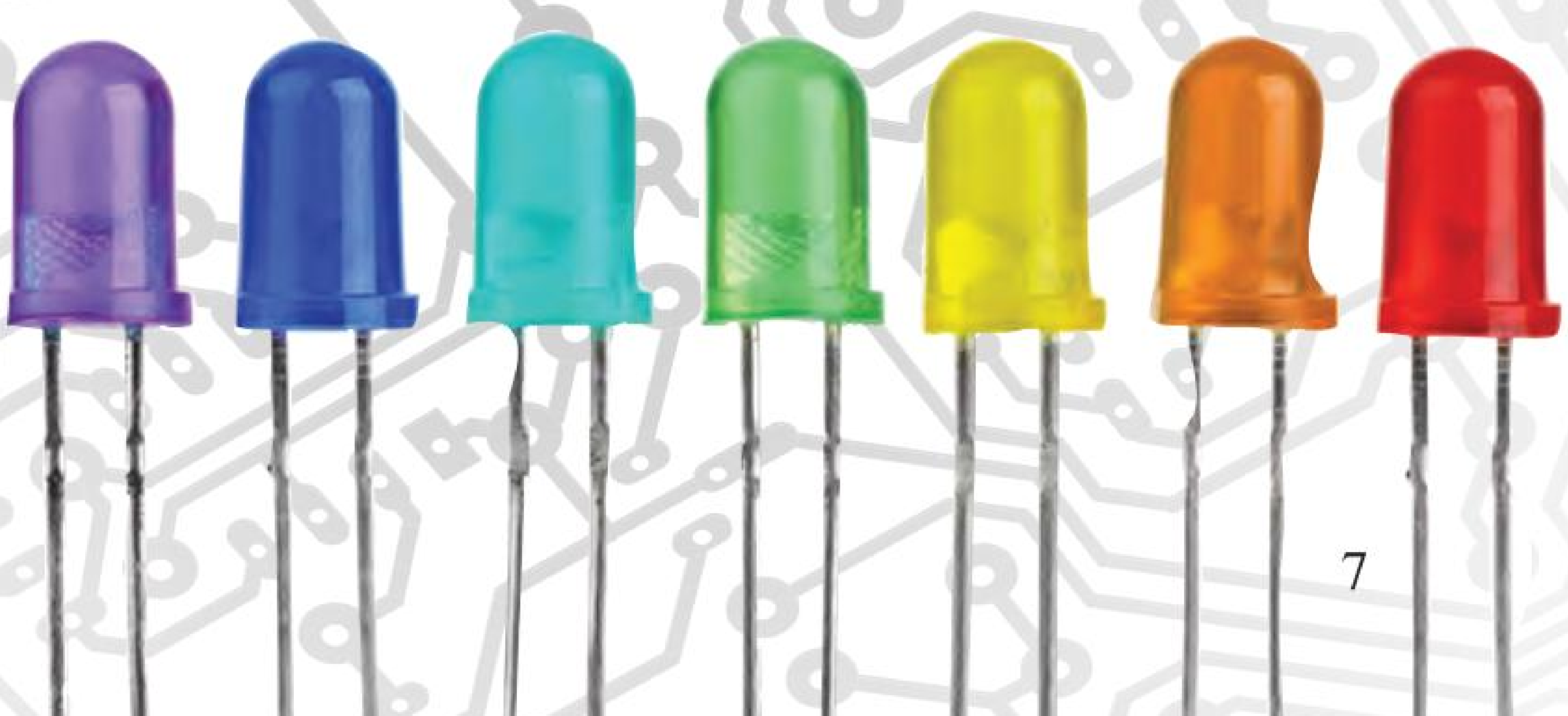
Max and his passion for flashing lights receives a boost in *Cool Beans*, as he explains his latest foray into smart tri-coloured LEDs. The combination of Arduino controllers and a neat little package called the NeoPixel is a gift for any light-based project designer.

Welcome Mike O'Keeffe

Last month, we had to bid a fond farewell to long-term *PIC n' Mix* columnist Mike Hibbett. As promised, this month, stepping into Mike's 32-bit-sized boots is another Mike – Mike O'Keeffe – who has completed Mike Hibbett's excellent *LPLC Oscilloscope* project. This is a most-promising start for our new *PIC n' Mix* scribe, and we wish him well as he settles into his new role.

Passive DI project

Our eyes were bigger than our belly page-wise this month, so our apologies for the lack of the promised *Passive Direct Injection Box* – we just couldn't squeeze it in! Fear not, it is a good project and we will include it in an upcoming issue.



NEWS

A roundup of the latest Everyday News from the world of electronics

Cinema innovates with 4K – report by Barry Fox

Slowly but very surely, digital projection has become the norm in cinemas. Only a few diehards still think they prefer 35mm or 70mm film, and occasionally a few specialist cinemas oblige with a screening. Unless the print is brand new, the dirt, scratches, jitter and flicker serve as a useful reminder of how much better digital projection is for everyday cinema.

Widescreen weekend

For serious film buffs, the National Media Museum, in Bradford, is holding a Widescreen Weekend (15-18 October). Bradford is one of only three venues in the world capable of screening 70mm widescreen, three-strip Cinerama and 35mm film formats under one roof. The weekend will celebrate the 60th anniversary of the Todd-AO* 70mm widescreen process with 'Oklahoma!' (1955), as well as staging a 70mm screening of Stanley Kubrick's '2001: A Space Odyssey' (1968) and a three-projector Cinerama screening of 'How the West Was Won' (1962) from an authentic Technicolor print.

4K digital cinemas

Back in the real world, cinema chain Vue International recently equipped all its 83 screens in the UK and Ireland with Sony Digital Cinema 4K projection systems, and is using them to screen 'live' plays from the UK's National Theatre, shot and post-produced in 4K.

Finding new markets

After 4K tests with the stage production of 'War Horse' in February 2014, the first end-to-end 4K production was 'Behind the Beautiful Forevers'. The play, by David Hare and based on Pulitzer Prize-winner Katherine Boo's book on slum life in Mumbai, India, was captured in performance in March 2015 with six Sony F55 cameras.

At a press preview screening in central London, Kevin Styles, UK & Ireland MD, Vue Entertainment, told how 'Event Cinema', theatre and sport on a big screen, is now Vue's 'fastest growing business, doubling year on year'.

Rufus Norris, of the National Theatre, said that screened theatre had doubled their audiences since the first 2K productions in 2009.

Colour space and quality

Sony says the colour space captured with the Sony F55s is far in excess of the Rec.709 standard for HD TV. The material is mastered in UHD (3840x2160 resolution) and a 4K DCP (digital cinema package) file created for cinema projection (4096x2160)

The 4K image clarity is certainly impressive, and the experience rewarding and enjoyable. But it is very different from attending the National Theatre's venues on the South Bank of the Thames in London – and not just because few people can afford seats that are close enough to the stage to compare with a screen experience. The use of close-ups and focus-pulling between characters is quite unlike live theatre, even as seen from the front row of the stalls, especially as 4K resolution heightens the defocussing effect.

On a non-technical level, the second half of the play at the press screening was more captivating than the first, simply because Vue had given everyone at the preview event a bag of noisy popcorn. Anyone munching loudly in a South Bank theatre would be quickly shushed and then ejected.

Fast charging

New low-cost smartphones from Motorola boast a feature called 'Turbopower', claimed to make them 'the world's fastest charging smartphones', with 10 hours of power stored from a mains charger in just 15 minutes. As Motorola demonstrated, in a comparative test staged at the London end of a teleconference link with Brazil, Turbopower far exceeds the charging speed of a Samsung Galaxy S6.

Operation

Finding anyone at the London event who could explain how the technology works, without risk of damage to the lithium ion batteries – or even fire caused by overheating – proved surprisingly difficult.

But persistence – and a magnifying glass on the microscopically small print on the charger itself – revealed that to use the feature the user must purchase an extra charger (made in China) which has a standard micro USB plug, but stepped DC output. Step 1 sends the normal 5V to the phone, but at high current (2.85A). Step 2, called 'Turbo 1', sends 9V (also at 2.85A), and Step 3, called 'Turbo 2', sends 12V at 2.15A.

Fast-charge-phone models

When connected to a conventional phone, the charger defaults to safe 5V working. However, when connected to a new Moto X Style or X Play phone, the charger handshakes with the phone via the USB data wires, and sends 12V. After partial charging is complete the chargers step down to 9V. Moto X Play will be available from the end of August and will cost £299; Moto X Style follows later this year with pricing as yet undecided. Details of availability and pricing for the Turbo chargers were unavailable.

Antex TCS230 launched

ANTEX
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350 °C

Digital temperature control at the touch of a button from Antex's new TCS230 soldering iron

New Antex digital soldering iron delivers the best of both worlds – Alan Winstanley reports

Every electronics constructor has an electric soldering iron in their toolbox, and many discerning enthusiasts choose a variable-temperature soldering station that offers more flexibility to cope with a wider variety of tasks. Soldering iron manufacturer Antex (Electronics) Ltd has launched a brand new concept in soldering irons that offers users the best of both worlds. The new Antex TCS230 is a mains-powered pencil-style soldering iron with a digital temperature control built into a new ergonomically shaped handle, which totally eliminates the need for a bulky bench-top station.

A backlit digital LCD temperature display and twin push buttons within the handle enable temperatures over the range 200 to 450°C (392 to 842°F) to be programmed directly into the iron at the touch of a button. The 230V, 50W rating is ample for all routine electronic and maintenance tasks, including lead-free soldering. A high-grade ceramic element offers very high insulation properties with leakage claimed as low as 5µA, making the Antex TCS230 suitable for working with sensitive microelectronics. Included as standard is their new-style slide-on, tool-free 2.3mm nickel-plated bit that makes the TCS230 ideal for general purpose use, and a full range of nine bits from a pinpoint 0.12mm to 6mm is available from Antex. A silicone mains cable protects against accidental damage, making the iron ideal for student or trainee use, and Antex offers a choice of UK or European mains plugs. A matching compact bench stand with sponge (the ST6A) is also available.

The new digital soldering iron will be the ideal solution for hobbyists and professionals seeking a powerful space-saving soldering iron with variable temperature and LCD display. The Antex TCS230 retails at £69.99 and is available from Antex and their distributors. Contact Antex (Electronics) Ltd, Westbridge Industrial Estate, Tavistock, Devon PL19 8DE, UK. Tel: +44 (0)1822 613565. Online, visit: **www.antex.co.uk** or email **sales@antex.co.uk** for more information.

Wireless charging for metal-body devices

Mobile phones that have metal bodies will soon be able to wirelessly charge themselves thanks to a new Qualcomm technology.

Wireless charging technologies that use induction heat metal, making them incompatible with any metal cases. Qualcomm's approach uses magnetic resonance, which is much more tolerant of metal items. It creates a charge over a small three-dimensional space so that coins, keys and other metal items in that space are unaffected by the charge.

It taps into a technology known as 'WiPower', which conforms to a standard called 'Rezence'. WiPower's magnetic resonance can operate at a frequency better able to handle metal objects without affecting the charging process. With this latest innovation, Qualcomm's WiPower can charge an entire device made of metal. According to Qualcomm, WiPower can charge devices that require as much as 22W at speeds that match or exceed those of other wireless charging technologies.

Braille smartwatch

A Korean company has created 'Dot' – the first smartwatch for the visually impaired. Dot has four sets of six dots, which raise to produce four braille characters at a time. New characters can rise at speeds ranging from 1-100 per second, which should suit all skill levels. See: **http://fingerson.strikingly.com**

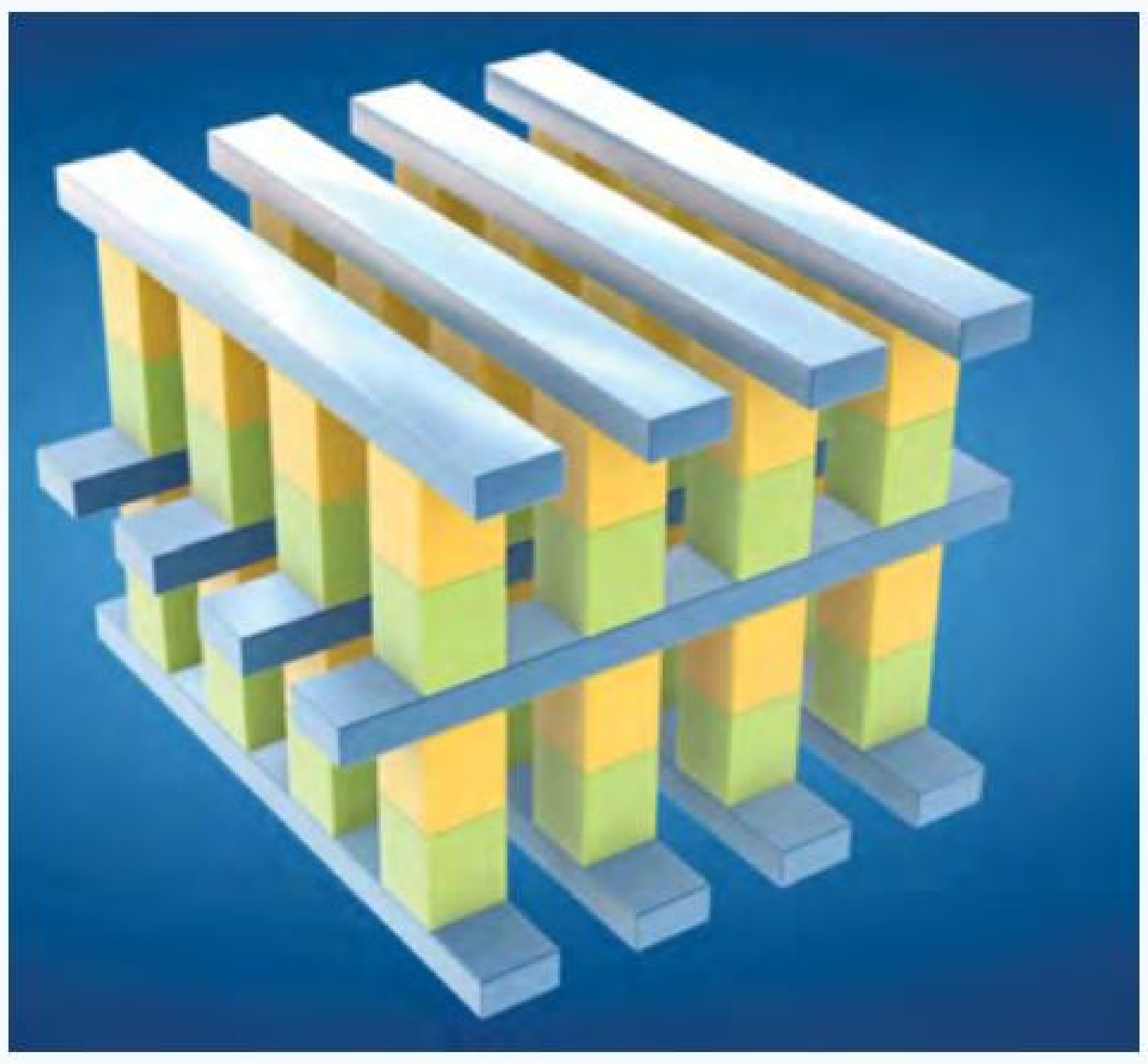
Breakthrough memory from Intel

Intel and Micron Technology have unveiled their new 3D XPoint technology, a non-volatile memory that has the potential to revolutionise any device, application or service that benefits from fast access to large sets of data. Now in production, 3D XPoint technology is a major breakthrough in memory process technology and the first new memory category since the introduction of NAND flash in 1989.

XPoint is up to 1,000 times faster and has up to 1,000 times greater endurance than NAND, and is 10 times denser than conventional memory.

The performance benefits of 3D XPoint technology could also enhance the PC experience, allowing consumers to enjoy faster interactive social media and collaboration as well as more immersive gaming experiences. The non-volatile nature of the technology also makes it a great choice for a variety of low-latency storage applications since data is not erased when the device is powered off.

The innovative, transistor-less cross point architecture creates a three-dimensional checkerboard where memory cells sit at the intersection of word lines and bit lines, allowing the cells to be addressed individually. As a result, data can be written and read in small sizes, leading to faster and more efficient read/write processes.



Intel's three-dimensional memory architecture delivers up to 1000-time speed increase

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TechnoTalk

Lateral thought

Mark Nelson

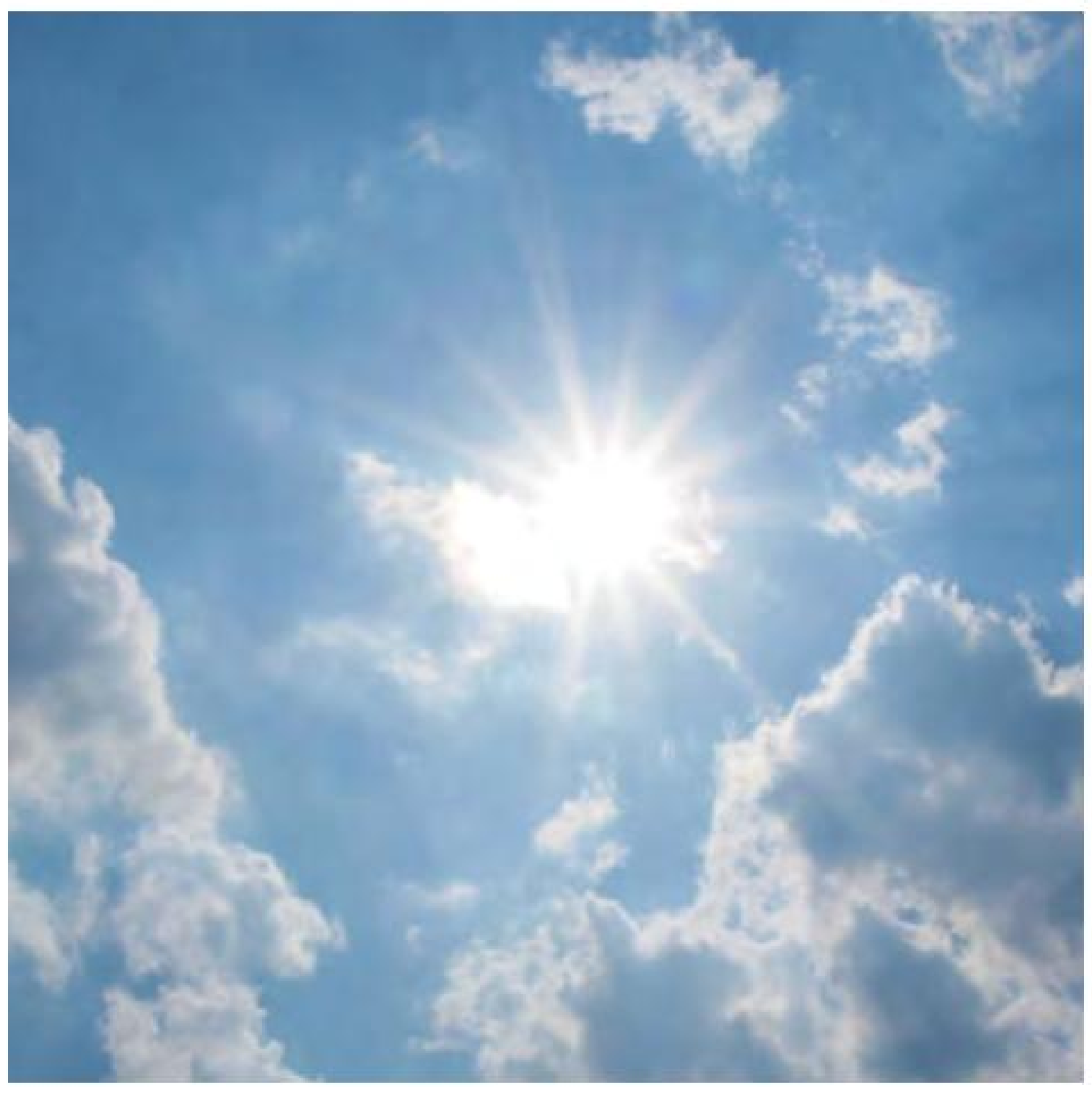
Solar power, one of the cleanest sources of renewable energy, may be poised to take a fundamentally new direction. Generating just volts may not be the best way of turning sunlight into power. This and other radical new trends in solar electronics are discussed by Mark Nelson.

HAVE YOU EVER STOPPED YOUR car to ask for directions, and received the following reply? 'If I were going there, I certainly wouldn't start from here...' Not very helpful, but it displays a refreshingly alternative line of thought. Precisely the same kind of lateral thinking applies to a remarkable new development in solar power – outputting fuel rather than electricity. The fuel is hydrogen gas that can be employed in the chemical industry or combusted in fuel cells – in cars for example – to drive engines. The principle has been demonstrated in working prototype form in the Netherlands, where it was devised by researchers at Eindhoven University of Technology. Solar fuels are a hugely promising replacement for fuels that pollute the environment, and if the process can be operated economically, the benefits are potentially enormous. So how does it work?

Nannoo, nannoo!

Remember this catchphrase from 'Mork and Mindy'? No? Nor do I, because I never watched the programme. It's no longer relevant anyway, because the new deal is 'nano, nano' – 'nano', as in nanowire. Nanowires, about 10 to 500nm (nanometres) in diameter – where 1nm is one thousandth of a micrometre, or a millionth of a millimetre – came up last month in connection with LEDs. Nanowires can also produce fuel from sunlight, by 'splitting' liquid water into its constituent hydrogen and oxygen parts using the electricity that is generated photovoltaically. You might assume that there's nothing special about the electrolysis process, but actually that's the clever part. To produce hydrogen gas from water, the solar panels use the material gallium phosphide (GaP), specifically in the form of very small nanowires. Using nanowires for this task is novel and helps to boost the yield by a factor of ten – and using ten-thousand times less precious material.

GaP has good electrical properties, but its drawback is that it cannot easily absorb light when a large flat surface is used in GaP solar cells. The researchers overcame this problem by making a grid of very small GaP nanowires, measuring 500nm long and 90nm thick. This immediately boosted the yield of hydrogen by a factor of ten to 2.9%. The university's professor Erik Bakkers explains the breakthrough is not simply about the yield, conceding there is still plenty of scope for improvement. But, he stresses: 'For the nanowires we needed ten-thousand times less of the precious GaP material than for cells with a flat surface. That makes nanowire cells potentially a great deal cheaper. In addition, GaP is also able to extract oxygen from the water – so you then actually have a fuel cell in which you can temporarily store your solar energy. In short, for a solar fuel's future we cannot ignore gallium phosphide any longer.'



Choose your own colours for LEDs

Graphene, an atomically thin and perfectly crystalline form of carbon, was featured in last month's article. This time, the spotlight is on phosphorene, a layer of phosphorus also one atom thick. The similarity ends there, however, because unlike graphene, phosphorene is a semiconductor, just like silicon, which is the basis of almost all of today's electronic technology. In Australia, scientists studying thin layers of phosphorus have discovered surprising properties that could open the door to ultrathin and ultralight solar cells and LEDs. It also shows very promising light-emission properties, states lead researcher Dr Yuerui (Larry) Lu, from the Australian National University (ANU).

'Because phosphorene is so slender and wispy, it opens possibilities for making lots of interesting devices,' he says. As well as creating semiconductors much thinner and lighter than silicon, phosphorene has light-emission properties that vary widely with the thickness of the layers, which enables much more flexibility for manufacturing. This property has never been reported before in any other material, he told the journal *Light: Science and Applications*. 'By changing the number of layers we can tightly control the band gap, which determines the material's properties, such as the colour of LED it would make. Under the microscope you can see quite clearly the different colours of the sample, which tells you how many layers are there,' concludes Dr Lu.

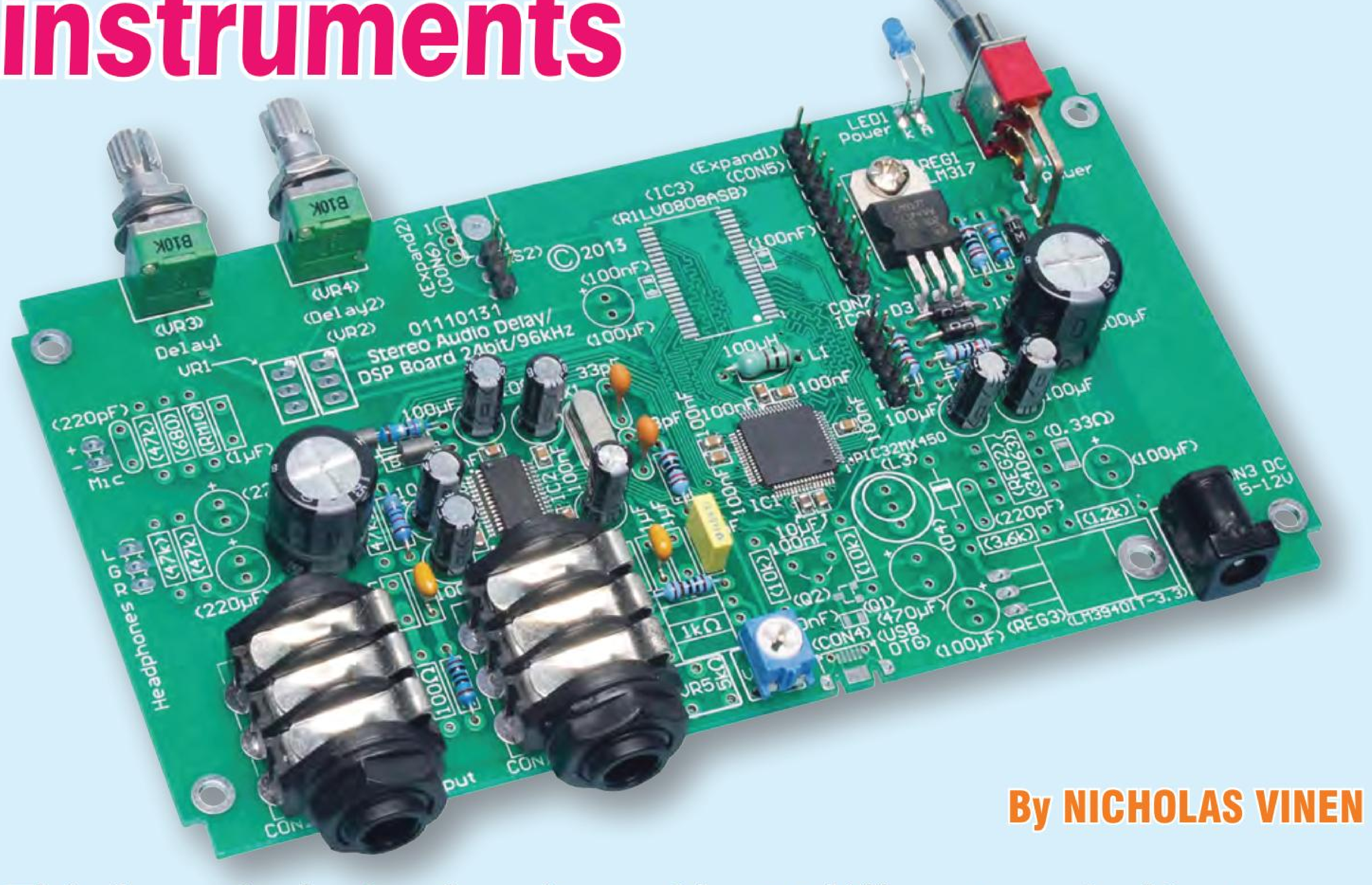
Solar Paper is here already

Dr Lu's team is still working on his ultra-thin solar cells, but a solar design start-up company in South Korea plans to have them on sale by the time you read this article – allegedly, anyway. As the **techinasia.com** website reports, a couple of green entrepreneurs plan to make solar power accessible to smartphone users, and since January of this year they have been working on making a portable solar charger that's as thin, powerful and aesthetically appealing as possible.

Called 'Solar Paper', the product fits snugly between the pages of a notebook and weighs just four ounces. In sunshine it can reliably charge your smartphone in about 2.5 hours, just like a wall charger. The basic model supplies 5W and if you need more power to revitalise a tablet or to charge on a cloudy day, you can click additional panels into place using their built-in magnets. Outdoorsy types can clip it to a backpack with carabiner clips supplied and charge while hiking. Urban adventurers can easily slip the panel into a jacket pocket or notebook. You can even see how the panel performs under different weather conditions, using a built-in LCD ammeter that shows exactly how much current is flowing into your device at any given time.

Retail price for the basic model has been set at around £77 and Solar Paper has its own web page at **https://www.kickstarter.com/projects/1398120161/solar-paper-the-worlds-thinnest-and-lightest-solar** where you can see how chic it looks, discover its ingenious features and place orders. Alternatively, go to **YouTube.com** and enter 'Solar Paper' (with the quote marks) to watch no fewer than ten videos on this imaginative device.

Digital effects processor for guitars and musical instruments



By NICHOLAS VINEN

This deceptively simple unit provides 10 different musical instrument effects, including echo, reverb, tremolo, fuzz, compression, flanging and phasing. Each effect is adjustable and can be defeated with a foot pedal switch. It's designed for use with electric guitars but will work with other instruments and vocals too.

WANT TO SPICE up your guitar performances? *Build this Digital Effects Processor* into a guitar amplifier and you will get many different effects to play with, without needing to lug around and wire up many different effects pedals. It can apply the majority of common effects to a line-level signal, and you can adjust them to suit your needs.

We can't promise that this will replace all your effects units but it certainly gives a lot of different options, which suit a variety of instruments, performers and musical styles. The idea is to build it into a guitar amplifier by connecting it between the preamplifier and amplifier sections. It can be powered directly from an amplifier supply rail, assuming a suitable DC voltage is available, or the supply rail can be derived, creating one very convenient package!

But it is not just intended for use with guitars. It is suitable for use with a large variety of other musical instruments, whether they are keyboards or instruments with pickups. And they can be used to enhance vocals as well.

Basically, if you want to add some pizazz to your performance, this *Digital Effects Processor* is a great way to do it.

Digital effects

The available effects are shown in Table 1. For each effect, there are two parameters that can be set using potentiometers VR3 and VR4. Those parameters are also listed in the table and described in the list of effects below.

Note that when one of the enabled effects causes a reduction in signal level (eg, echo or reverb), the level for all effects is reduced, as well as the level when no effect is selected, in order to prevent noticeable changes in signal when switching between them.

The available effects are as follows:

- **Echo:** delays / attenuates the incoming signal, then mixes it back in for the output. VR3 adjusts the delay between (nearly) 0ms and 1200ms with an exponential curve, to make it less sensitive at the shorter end, which is more useful. VR4 adjusts the attenuation level; at higher settings, the echo is louder. Note: as the echo becomes louder, the original signal becomes quieter to prevent overload.
- **Reverb:** the same as echo, except that many extra short echoes are added to simulate reflections from multiple hard surfaces in close proximity.
- **Tremolo:** the output volume is modulated by a sinusoidal waveform. VR3 adjusts the amount of modulation (ie, 'depth') while VR4 changes the frequency.
- **Vibrato:** the output frequency is modulated by a sinusoidal waveform. VR3 adjusts the amount of modulation (ie, 'depth') while VR4 changes the frequency. Note that this is performed by slightly speeding up and slowing down the audio signal, although the change in delay that this causes should be imperceptible.
- **Overdrive:** this provides adjustable clipping for the signal. VR3 adjusts the gain applied to the signal and once the amplitude is high enough, it clips. VR4 adjusts how progressively the clipping occurs; at minimum setting it is hard, resulting in a square wave, while at higher settings for VR4, the clipping is more progressive and the waveform becomes rounded.
- **Fuzz:** the same as overdrive, except that the gain is applied asymmetrically, in order to inject extra distortion into the signal.
- **Compression:** the gain is slowly increased until the output reaches 90% of maximum. If the output exceeds 90%, the gain is decreased. VR3 sets the rate of increase while VR4 sets the rate of decrease. The scale for VR4 is different for VR3 as the rate of decrease is normally much higher.
- **Noise gate:** similar to but not quite the opposite of compression. When the input signal is below the threshold, there is no output. When the input goes above the threshold, it is sent to the output. VR3 adjusts the threshold while VR4 adjusts the hysteresis, to prevent the output from fluctuating on and off with a signal near the threshold.
- **Flanger:** this mixes the input signal with a version of the signal that has slight vibrato applied, causing a distinctive 'comb filter' Doppler effect.
- **Phaser:** similar to flanger but mixes the signal with a version that has a modulated phase shift, causing a 'rippling' effect which makes the sound seem artificial.

Features & Specifications

- 10 effects to choose from: echo, reverb, tremolo, vibrato, overdrive, fuzz, compression, noise gate, flanger and phaser
- Each effect has two adjustable parameters
- Maximum echo/reverb delay: 1.2 seconds
- Four-position switch selects between three effects and no effect with seamless transitions
- Optional defeat switch (eg, foot pedal)
- Low noise and distortion: THD+N typically <0.02%, signal-to-noise ratio >76dB
- Two power supply options: 3.5-6V DC / 7.5-12V DC; current drain 60-80mA
- Optimal input signal range: 0.5-2V RMS
- Line output signal: typically 1V RMS
- Input impedance: 4-6kΩ
- Optional headphone output
- Optional microphone preamplifier

Modes

The effects are selected using a 4-position rotary switch (a slide switch could also be used). The second position selects no effect and the other three positions can each select one of the 10 effects listed above. We've made the second position the 'off' position to make it easy to switch between two commonly used effects and none.

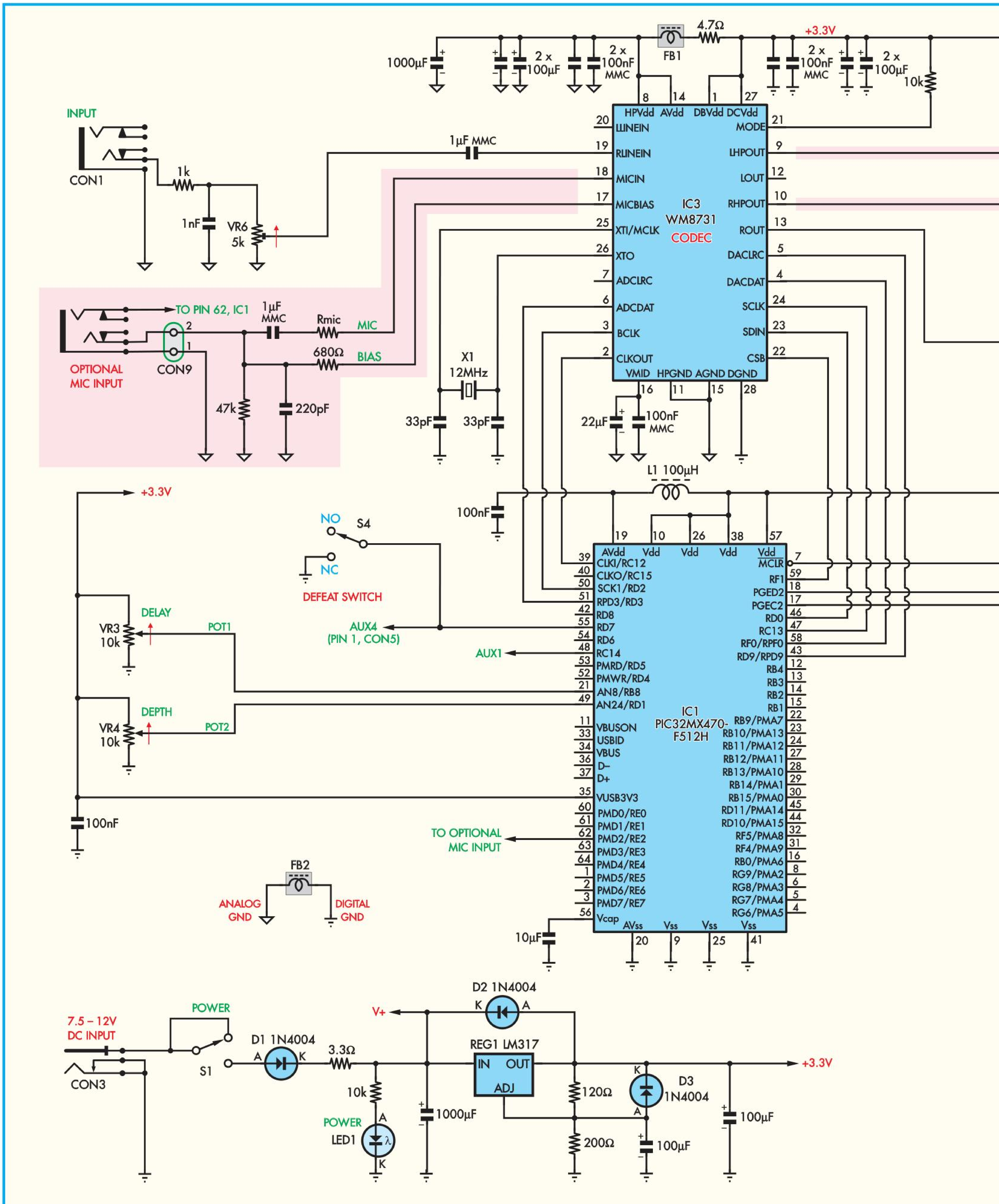
A pushbutton is used to change which effect is selected by each switch position and is also used when adjusting the two knobs, so that different settings can be used for each effect and they don't have to be reset each time a different effect is used.

The same effect can be used with different settings for each switch position. If you don't need three effects, the switch can be limited to fewer positions. There is also a simple fall-back mode available; if the pushbutton is permanently shorted out (say, on the PCB) then the three effects selected by the rotary switch are always echo, reverb and tremolo, and VR3 and VR4 can be adjusted at any time.

A defeat switch (eg, a foot switch) can be added and this has the same effect as switching the rotary switch to the 'off' position as long as it is held down. Or you can wire it the other way around, so that effects are only applied when the switch is held down.

Table 1: Effects controls

#	Effect	VR3	VR4
1	Echo	Echo delay	Echo fall-off
2	Reverb	Reverb delay	Reverb fall-off
3	Tremolo	Amplitude	Rate
4	Vibrato	Amplitude	Rate
5	Overdrive	Gain	Softness
6	Fuzz	Gain	Softness
7	Compression	Attack	Decay
8	Noise Gate	Threshold	Hysteresis
9	Flanger	Amplitude	Rate
10	Phaser	Amplitude	Rate


1000µF
2 x 100µF
2 x 100nF MMC
FB1
4.7Ω
+3.3V
2 x 100nF MMC
2 x 100µF
10k
INPUT
CON1
1k
1nF
VR6 5k
1µF MMC
IC3 WM8731
CODEC
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47k
220pF
X1 12MHz
33pF
33pF
22µF
100nF MMC
L1 100µH
100nF
+3.3V
NO
S4
NC
DEFEAT SWITCH
DELAY
VR3 10k
POT1
AUX4 (PIN 1, CON5)
AUX1
DEPTH
VR4 10k
POT2
100nF
TO OPTIONAL MIC INPUT
FB2
ANALOG GND
DIGITAL GND
10µF
IC1 PIC32MX470-F512H
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CLKI/RC12
CLKO/RC15
SCK1/RD2
RPD3/RD3
RD8
RD7
RD6
RC14
PMRD/RD5
PMWR/RD4
AN8/RB8
AN24/RD1
VBUSON
USBID
VBUS
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VUSB3V3
PMD0/RE0
PMD1/RE1
PMD2/RE2
PMD3/RE3
PMD4/RE4
PMD5/RE5
PMD6/RE6
PMD7/RE7
Vcap
AVss Vss Vss Vss
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RD0
RC13
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S1
D1 1N4004
3.3Ω
10k
POWER
LED1
V+
D2 1N4004
REG1 LM317
IN OUT
ADJ
1000µF
120Ω
200Ω
D3 1N4004
100µF
100µF
+3.3V

DIGITAL EFFECTS PROCESSOR

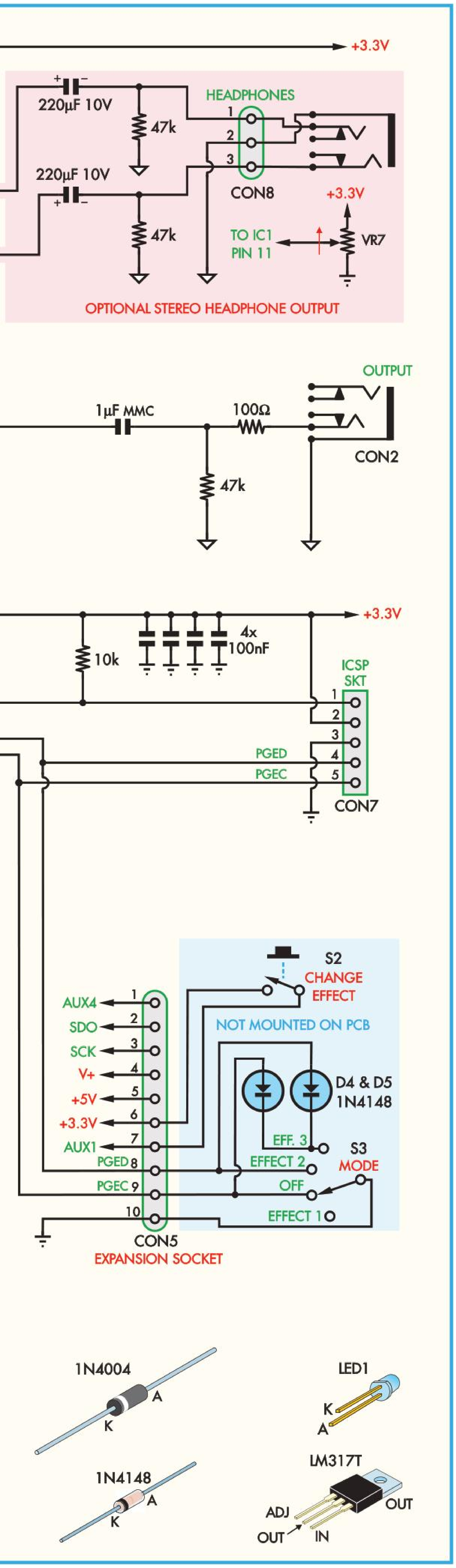

Fig.1: the basic *Digital Effects Processor* circuit. The incoming audio analogue signal at CON1 is digitised by CODEC IC2 and then fed to IC1 where it is processed and then sent back across the same digital audio bus to IC2. A DAC in IC3 then converts it back into an analogue signal, which is fed to the output (CON2).

Options

This *Digital Effects Processor* uses the same hardware as the *Stereo Echo & Reverb Unit* (March 2015) and the *Dual Channel Audio Delay* (February 2015). However, we have removed a number of components which aren't needed. For example, most musical instruments are not stereo, so components are only fitted for one channel (and indeed, the software only supports one channel).

As with those earlier designs, it is possible to add extra components to provide a microphone input or stereo headphone output. The processed mono signal is sent simultaneously to both headphone output channels.

The headphone output could be useful for monitoring purposes. It's up to you whether you want to install the few extra components required, which are shown in the circuit diagram at upper-right and on the overlay diagram, labelled in green.

The microphone input is less useful as its signal-to-noise ratio is only average. For a musical performance, you would be better off using an external microphone preamplifier such as our *High-Performance Microphone Preamplifier* from the June 2012 issue, which can run from the same DC voltage source as the *Digital Effects Processor* unit.

Software

In adding these new effects to the software, we have made some other changes at the same time. By making it process only a mono signal, this doubles the maximum echo to 1.2 seconds without needing an external RAM chip. This is more than long enough for instrumental work and so we haven't even bothered to provide the option of extra RAM in the software.

We've also gone to some effort to make changes between effects and changes in effect settings 'seamless', so that clicks and pops are not generated during a performance, even if settings such as echo delay are adjusted live.

Circuit description

The circuit diagram of the *Digital Effects Processor* is shown in Fig.1. As stated earlier, this is a simplified version of the circuit for the *Stereo Echo & Reverberation Unit* from the March 2015 issue, with unnecessary components removed. That's why there are so many unconnected pins on IC1; those originally used for interfacing with the unused SRAM chip and USB socket are not connected to anything.

A line-level signal, from a guitar preamp or mic preamp is fed into CON1 (connector tip). RF signals that may have been picked up are rejected by a low-pass filter comprising a 1kΩ series resistor and 1nF capacitor to ground, while 5kΩ trimpot VR6 is used to reduce the level to no more than 1V RMS, the limit of what the CODEC can handle.

The signal is then AC-coupled to the right channel input of the CODEC (IC3) via a 1µF DC-blocking capacitor. A half-supply (~1.65V) DC bias for this input is provided by the IC itself

Alternatively, a microphone signal can be applied to a 3.5mm jack socket connected to pin header CON9 and this is coupled to IC3's microphone input pin (pin 18) via a 1µF capacitor and optional series resistor (Rmic) which reduces the amount of gain if fitted; otherwise it is linked out. IC3 can supply a bias current for electret microphones and this is fed via a 680Ω series resistor. The associated 220pF capacitor provides some RF filtering for the microphone signal.

The microphone input is selected when the RE2 input of IC1 (pin 62) is pulled low. This is wired to the microphone socket so that the sleeve of the mono jack plug shorts it to ground when it is inserted. When this line is open-circuit, the line input is the active input. If the microphone input is not needed, the components in the pink box at left do not need to be installed.

CODEC operation

Whichever signal is selected, it is digitised by IC3 with a sampling rate of around 40kHz, and the resulting PCM digital audio signal is transmitted to PIC32 microcontroller IC1 via an I^{2}S bus. This appears at pins 3, 5 and 6 of IC3, which are the serial bit clock, sample clock and serial data line respectively. These connect to the audio CODEC-compatible SPI peripheral in IC1.

IC1 reads the digital audio data from the CODEC, processes it to add the selected effect (depending on the mode) and also stores it within its 128KB

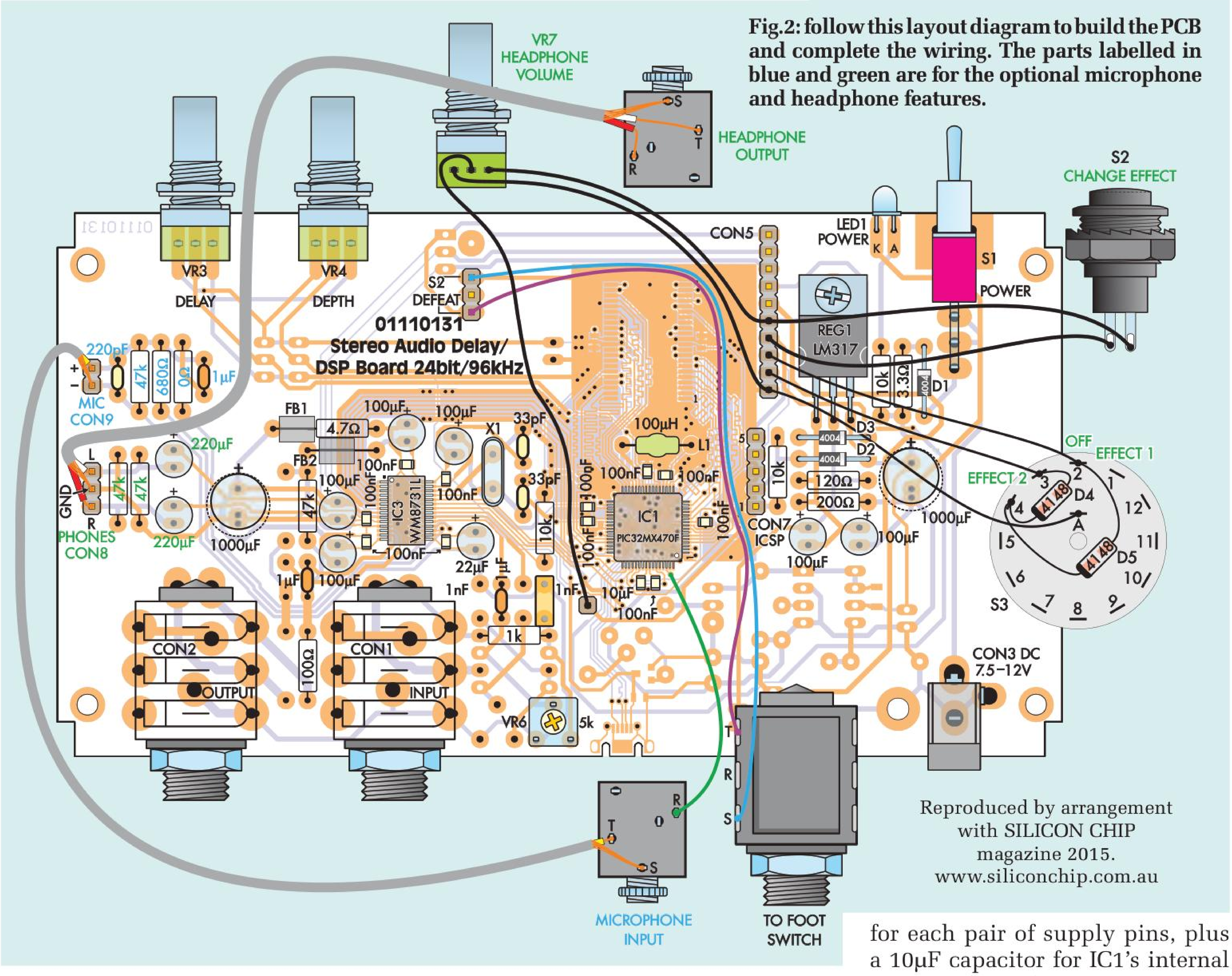
VR7
HEADPHONE
VOLUME
HEADPHONE
OUTPUT
S
T
R
S2
CHANGE EFFECT
VR3
DELAY
VR4
DEPTH
S2
DEFEAT
01110131
Stereo Audio Delay/
DSP Board 24bit/96kHz
CON5
LED1
POWER
K A
S1
POWER
REG1
LM317
10k
3.3Ω
4004
D1
D2
D3
220pF
47k
680Ω
0Ω
1μF
MIC
CON9
FB1
4.7Ω
100μF
100μF
33pF
X1
100μH
L1
FB2
100nF
100nF
100nF
100nF
IC3
WM8731L
IC1
PIC32MX470F
100nF
120Ω
200Ω
CON7
ICSP
100μF
100μF
1000μF
OFF
EFFECT 1
EFFECT 2
D4
D5
S3
1 2 3 4 5 6 7 8 9 10 11 12
A
GND
L
R
PHONES
CON8
220μF
220μF
1000μF
1μF
100μF
22μF
1nF
1nF
10μF
1k
10k
47k
100Ω
CON2
OUTPUT
CON1
INPUT
VR6
5k
CON3 DC
7.5–12V
R
T
S
MICROPHONE
INPUT
TO FOOT
SWITCH

Fig.2: follow this layout diagram to build the PCB and complete the wiring. The parts labelled in blue and green are for the optional microphone and headphone features.

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RAM, for the echo and reverb effects. Processed audio data is sent back over the same I^2S bus, this time to pin 4 of IC3 but timed using the same clock lines. The CODEC then converts this digital stream back to analogue audio data, which it transmits from its line out (pin 13) and headphone out (pins 9 and 10).

These signals are all AC-coupled to the respective output connectors, to remove the 1.65V DC bias, via a 1μF capacitor for the line output and 220μF capacitors for the headphone outputs. The reason the headphone output needs much larger capacitors is that the headphones will have a much lower impedance than the line input of other equipment; 8-600Ω for headphones compared to several kilohms for a line input.

The line output also includes a 100Ω series resistor, both to prevent cable capacitance from causing instability in the output drivers of IC3 and to protect IC3 against a shorted output.

IC3 also contains a digital volume control which adjusts the headphone amplifier output. If VR7 is fitted to the board, IC1 detects this and sends commands to IC3 to set the headphone volume depending on the voltage at VR7's wiper. If VR7 is not fitted, the headphone outputs are disabled and in that case, the other components in the pink box may be omitted.

Controls and power supply

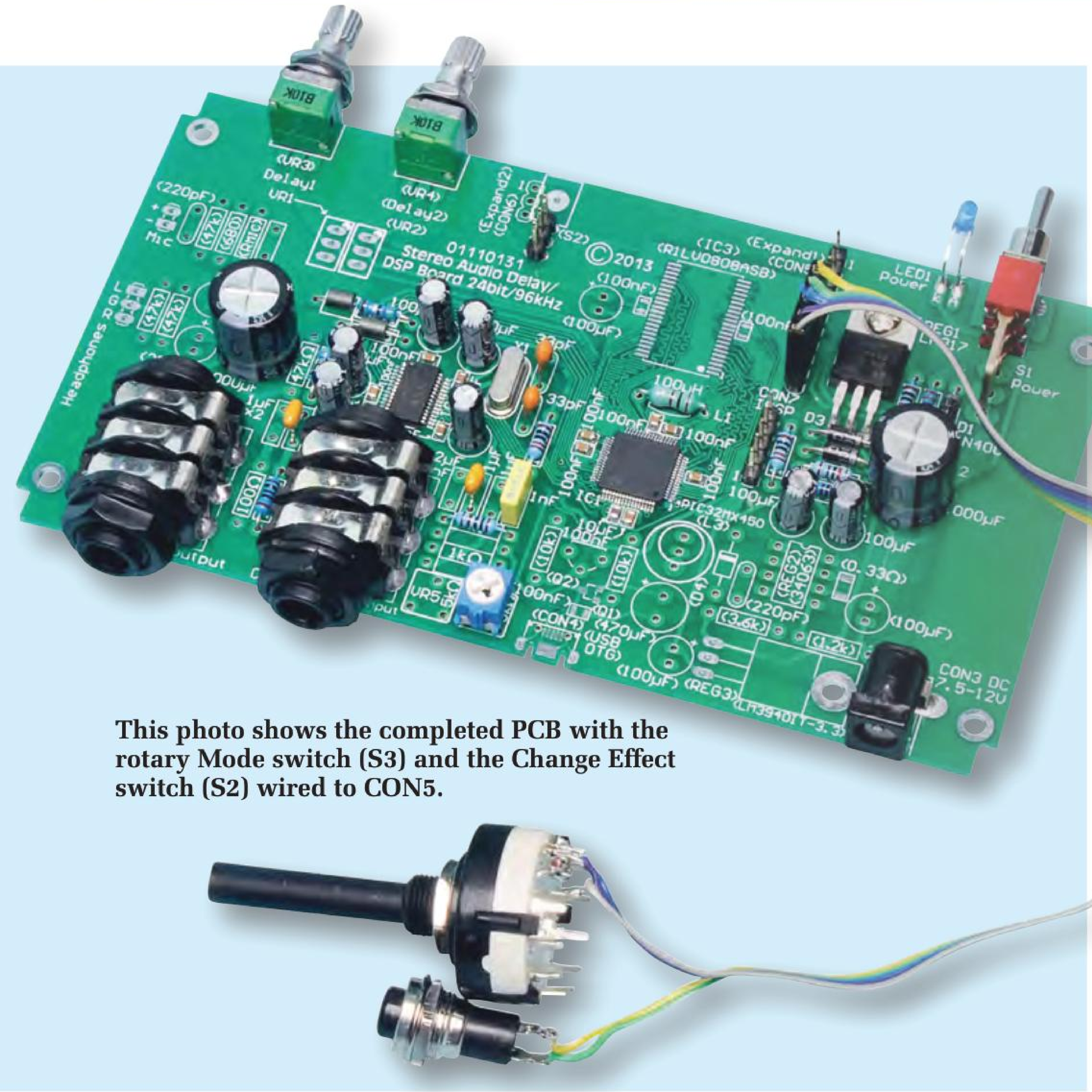
Pots VR3 and VR4 are used to change the effect parameters. These form voltage dividers across the 3.3V supply rail, and the wiper voltage is read by IC1 using its internal analogue-to-digital converter (ADC).

The power supply is quite simple. D1 provides reverse polarity protection while REG1 drops the incoming 7.5-12V rail to a regulated 3.3V, as required by IC1 and IC3. LED1 indicates when power is applied. IC1 and IC3 have 100nF bypass capacitors for each pair of supply pins, plus a 10μF capacitor for IC1's internal core regulator (on pin 56, Vcap).

CODEC IC3 also has 100μF bypass capacitors for each supply pin to ensure a low supply impedance and thus good performance. Its analogue and digital grounds are separated by ferrite bead FB2 to minimise digital noise coupled into the analogue ground, where it could otherwise reach signal paths. A 4.7Ω series resistor also provides audio-frequency low-pass filtering for the analogue supply, in conjunction with the 1200μF capacitance on the analogue supply rail (ie, 1000μF plus 2 × 100μF).

5V operation

As with the *Echo & Reverb* unit, you can change some components to operate the unit from a 5V supply, such as is available from a USB port. This arrangement is shown in Fig.3. Basically, REG1 and its associated components are deleted and an LM3940 low-dropout 3.3V linear regulator is substituted. This is necessary because the LM317 used for higher voltage supplies drops



This photo shows the completed PCB with the rotary Mode switch (S3) and the Change Effect switch (S2) wired to CON5.

too much voltage and can't operate from 5V. Also, D1 is replaced with a 1N5819 Schottky diode, which has a much lower forward voltage.

Construction

Fig.2 shows the PCB (available from the *EPE PCB Service*, code 01110131) parts layout. If building the 5V-powered version, refer also to Fig.4 for the necessary changes to fit the different regulator and Schottky diode (D1).

Start by fitting SMDs IC1 and IC3 (IC2 is left out). In each case, place the IC alongside its pads, right-side up and identify pin 1 (there should be a depression in one corner but magnification may be required to spot it). A pin 1 dot is also shown on the overlay diagram and PCB.

Apply a very small amount of solder to one of the corner pads. If you are right-handed, it's easiest to start with the top pad on the right side, or if left-handed, with the top pad on the left side. Avoid getting any solder on the adjacent pad.

Now, pick up the part with a fine-tipped pair of angled tweezers and while heating this pad, gently slide the IC into place. Check the part's alignment under a magnifying lamp. All the pins must be centred fairly accurately over their respective pads.

If they aren't, don't panic, it's just a matter of re-melting the solder on that one joint and carefully nudging the IC in the required direction, then re-inspecting it. It may take a few attempts to get it correct. Care and patience are a virtue here, the goal being to eventually get it properly aligned without spreading solder onto any more pins or pads and without heating the PCB or IC enough to damage them.

Once the part is in place, solder the diagonally opposite pin, then re-check the alignment under magnification as it may have moved slightly. If it has, you can reheat this second pad and gently twist the IC back into alignment. Once you're happy, proceed to solder the remaining pins without worrying too much about bridging them (it's hard to avoid). Remember to refresh that first pin you soldered.

Now spread a thin layer of flux paste along all the pins and gently press down on them with solder wick and a hot iron to suck up the excess solder. If done correctly, this will leave you with neatly soldered pins and no solder bridges. Go over all the pins once with the solder wick, then check under a magnifier for any remaining bridges. If there are any, add a dab of flux paste and go back over them with the solder wick.

With all the joints looking good, you can install the other SMD IC using the

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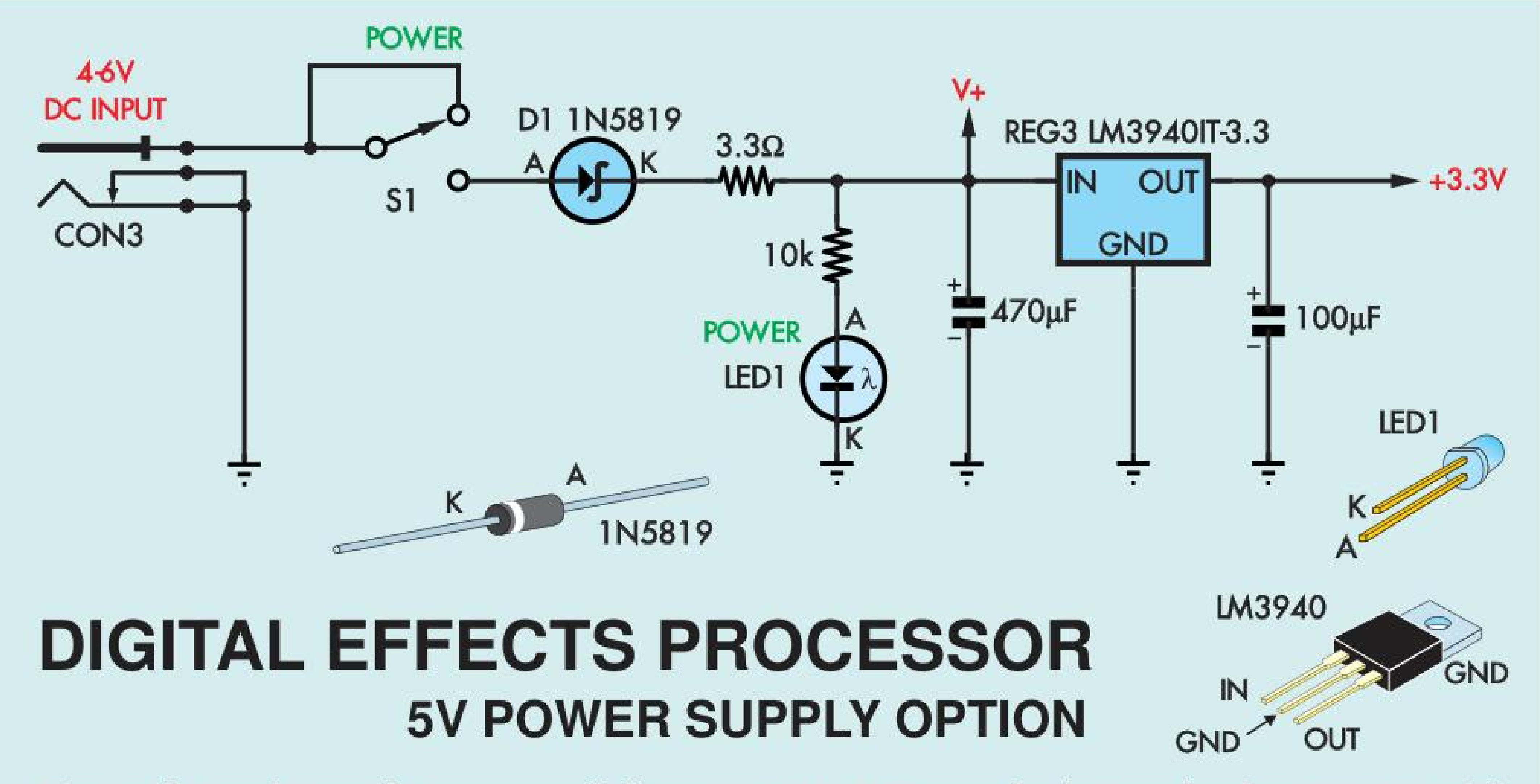


Fig.3: the unit can be powered from a 4-6V DC supply by replacing REG1 with an LM3940IT-3.3 low-dropout regulator and changing D1 to a 1N5819.

same procedure. Note that a hot-air gun/toaster oven and solder paste can also be used for these ICs.

Once you've checked that the ICs are all soldered properly, follow with the SMD ceramic capacitors, using a similar procedure; ie, add solder to one pad, heat and slide the part into place, then solder the other pad and refresh the initial joint. Don't get the 10µF capacitor mixed up with the others.

You do need to be careful to wait about 10 seconds after soldering one side of a capacitor before applying solder to the other side though. The capacitors are so small that the solder joint can remain molten for quite some time. If you try to solder the opposite pad too early, the capacitor will move out of alignment and it's frustrating to re-align capacitors when this happens, especially if solder has taken to the other pad too.

So take it slowly and be careful not to short any of the adjacent IC pins when soldering the pads; the capacitors have been placed quite close for performance reasons. A fine soldering iron tip will make this easier.

Through-hole parts

Proceed now with the low-profile components such as resistors and diodes – remember to slip a ferrite bead over the 4.7Ω resistor lead before soldering it in place. It's best to check each resistor value with a DMM before fitting it as the colour bands can be difficult to read. The diodes are all the same type and all have their cathode bands facing to the top or right edge of the board.

For FB2, slip another bead over a resistor lead off-cut and then solder it

Parts List

1 double-sided PCB, available from the *EPE PCB Service*, coded 01110131, 148 × 80mm
1 12MHz HC-49 crystal (X1)
1 100µH axial RF inductor (L1)
2 10kΩ 9mm horizontal potentiometer (VR3,VR4)
1 5kΩ mini horizontal trimpot (VR6)
2 6.35mm PCB-mount stereo switched jack sockets (CON1,CON2)
1 10-way pin header, 2.54mm pitch (CON5)
1 5-way pin header, 2.54mm pitch (CON7) (optional)
1 PCB-mount SPDT right-angle toggle switch (S1)
1 chassis-mount NO momentary pushbutton switch (S2)
1 4-position rotary or slide switch (S3)
1 3-way pin header, 2.54mm pitch (for S4)
1 foot switch with cable (S4, optional)
1 DC plugpack, 7.5-12V, 100mA+
1 PCB-mount switched DC socket to suit plugpack
2 4mm ferrite suppression beads
9 M3 × 6mm machine screws
1 M3 nut
4 tapped spacers
1 metal case (optional)
Light duty hook-up wire/ribbon cable

Semiconductors
1 PIC32MX470F512H-I/PT 32-bit microcontroller programmed with 0120914A.hex (IC1) (the software to programme your own PIC is available from our website www.epemag.com, alternatively pre-programmed PICs are available from www.siliconchip.com.au)
1 WM8731SEDS 24-bit 96kHz stereo CODEC (IC3)
3 1N4004 diodes (D1-D3)
2 1N4148 diodes (D4,D5)
1 LM317T adjustable regulator (REG1) (refer to text for parts required for 5V DC operation)
1 3mm blue LED (LED1)

Capacitors
2 1000µF 25V electrolytic
6 100µF 16V electrolytic
1 22µF 16V electrolytic
1 10µF 6.3V 0805 SMD ceramic
1 1µF 50V monolithic ceramic
11 100nF 6.3V 0805 SMD ceramic
1 1nF MKT
2 33pF ceramic disc

Resistors (0.25W, 1%)
1 47kΩ
5 10kΩ
1 1kΩ
1 200Ω
1 120Ω
1 100Ω
1 4.7Ω 0.5W 5%
1 3.3Ω 0.5W 5%

Extra parts for headphone output
1 panel-mount stereo jack socket
1 10kΩ linear potentiometer, panel mount (VR7)
1 small knob to suit
2 220µF 10V electrolytic capacitors
2 47kΩ 0.25W resistors
1 3-way pin header
1 100mm length 2-core shielded cable or 3-strand ribbon cable
1 100mm length 3-strand ribbon cable

Extra parts for microphone input
1 3.5mm panel-mount stereo jack socket
1 1µF multi-layer ceramic capacitor
1 220pF ceramic capacitor
1 47kΩ 0.25W resistor
1 680Ω 0.25W resistor
1 2-way pin header
1 length shielded cable
1 length light-duty hookup wire

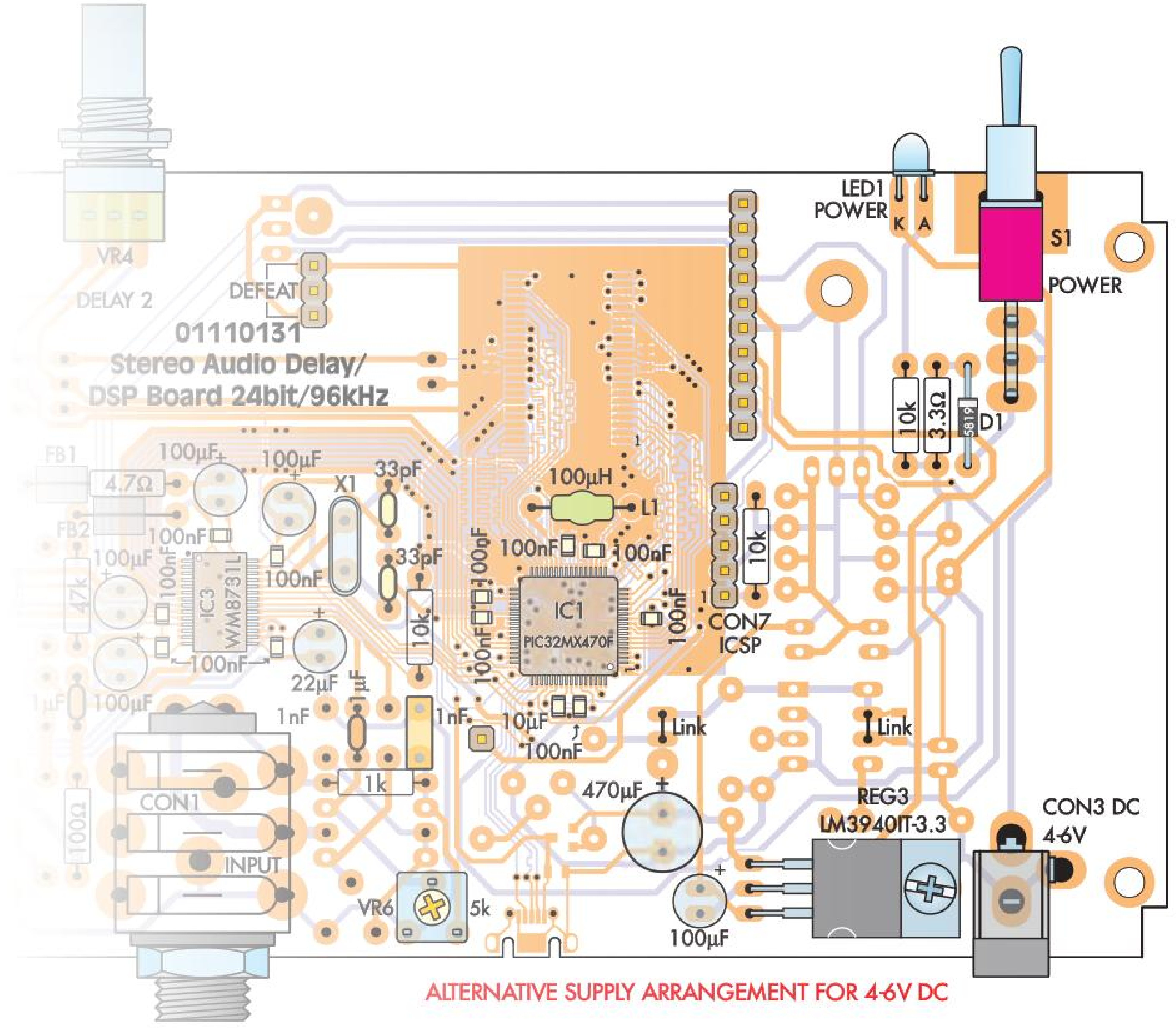


Fig.4: follow this PCB parts layout diagram to install the parts for the 5V power supply option. Note that D1 must be changed to a 1N5819 Schottky type.

to the board. You can also mount axial inductor L1 at this point. Then fit REG1 or REG3 (depending on supply voltage); bend its leads down about 6mm from its body, feed them through the holes, screw its tab to the PCB tightly and then solder and trim the leads.

Horizontal trimpot VR6 can go in next, followed by the ceramic capacitors (disc and monolithic multilayer) and then pin headers CON5 and CON7, plus the one for S4. You will also have to fit CON8/CON9 if you are using those optional features. Note that CON7 is not required if you have a pre-programmed microcontroller.

Now solder the DC socket in place, followed by VR3 and VR4. Note that you could mount all these components off-board (eg, chassis mount them) and run them back to the pads via flying leads, if that suits your application. This would be the way to fit the unit into a guitar amplifier, for example.

You can then fit crystal X1 and the electrolytic capacitors, of which there are three different values (four if using the headphone outputs). As usual, the longer lead is positive and this should go in the hole marked with the '+' symbol on the overlay, ie, towards the top edge of the board.

Next, fit power switch S1 and the power LED. The latter should have its lead bent at right angles 4mm from the base of the lens and then soldered so that the centre of the lens (and thus this short lead section) is 6.5mm above the top surface of the PCB. This aligns the centre LED with the centre of the switch. When bending the leads, pay attention to the 'A' and 'K' markings on the PCB – the longer (anode) lead of LED1 must be soldered to the anode pad.

S1 and the power LED could also be chassis-mounted if you wish.

The PCB assembly can now be completed by soldering jack sockets CON1 and CON2 in place. You will also need to wire up a rotary or slide switch (S3) and a momentary pushbutton switch (S2) to a pin header socket, as shown in Fig.2.

If using the foot (defeat) switch, headphone or microphone options, wire them up too. We've shown the foot switch connected via a 3.5mm phono socket, but you could use a 6.35mm socket or some other connector instead.

Checking it out

If you used a blank PIC32 chip, program it now. The circuit can be powered from a PICkit3 at 3.3V. In fact, the whole unit will operate normally from this supply, so you can test the audio signal path immediately after programming the chip. Use the firmware for the *Digital Effects Processor* which is available from our website **www.epemag.com** named '0120914A.hex'.

If you don't have a PICkit3, you will need to power the unit from a DC plugpack for testing. In this case, connect a voltmeter across the 3.3Ω resistor next to D1. Small alligator clip leads (or other test probe clips) are very useful for this purpose as you can switch the unit on while watching the meter reading and switch it off quickly should the voltage across this resistor rise too high.

Expect a reading in the range of 0.2-0.3V, depending on the exact resistor value and how you have configured the unit. Much less than 0.2V indicates that there is an open circuit somewhere, while much more than 0.3V indicates a likely short circuit. If the reading is outside the expected range, switch off immediately and check for faults.

The most likely faults would be pins on the SMD chips bridged to an adjacent pin or not properly soldered to the PCB pad, followed by incorrect device orientation (primarily ICs, diodes and electrolytic capacitors) or poor/bridged through-hole solder joints.

Assuming all is OK, connect S3, set it to position 2 (no effect) and feed a signal into the input; if it's a stereo plug, the left channel will be shorted out. Connect the output to some sort of amplifier. You should hear clear, undistorted audio with no effects. You can then try out the effects to check that they operate as expected.

Using effects

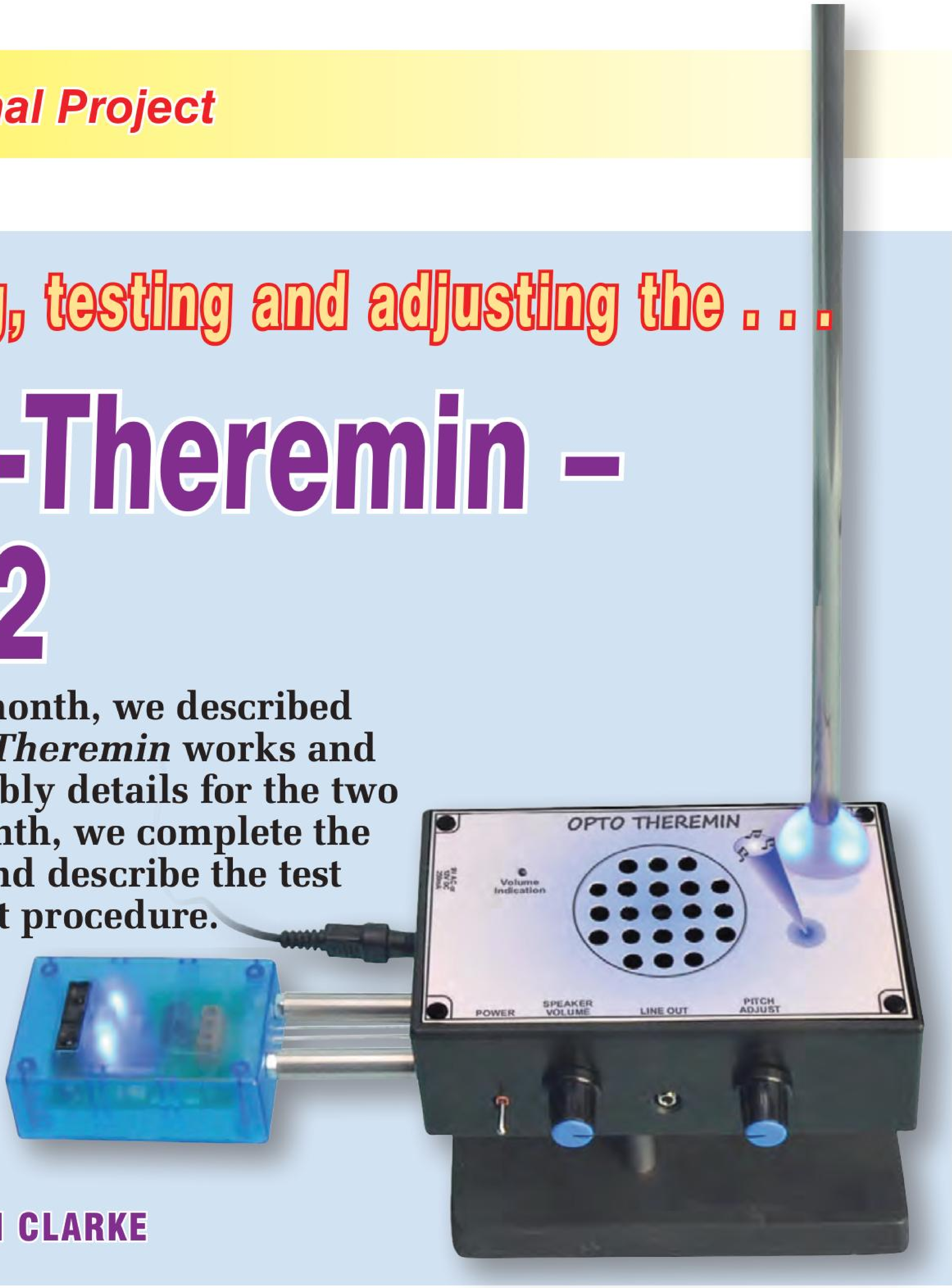
Initially, the effect for switch position #1 is echo, position #3 is reverb and position #4 is tremolo, so you can easily try these out. To adjust the parameters, hold down S2 and then rotate VR3 and/or VR4. Once you release S2, turning VR3 and VR4 will have no effect, so you can't accidentally change the settings.

To assign a different effect to one of these switch positions, select that position and then give S2 a brief press without turning either VR3 or VR4. The unit will switch to the next effect and will emit a series of 'pips' from the audio output; one pip for effect #1 (echo), two for effect #2 (reverb) and so on. If you press S2 when effect #10 (phaser) is selected, it will switch back to #1 (echo).

The settings are remembered even when power is removed; they're stored in Flash memory. If you press S2 in order to adjust VR3/VR4 and then decide against it, hold S2 down for a short period before releasing it to prevent an undesired change in the selected effect. Any press longer than about half a second will not cause the selected effect to change.

Completing, testing and adjusting the . . .

Opto-Theremin – Part 2



In Part 1 last month, we described how the *Opto-Theremin* works and gave the assembly details for the two PCBs. This month, we complete the construction and describe the test and adjustment procedure.

By JOHN CLARKE

THE OPTO-THEREMIN'S main PCB is housed in a black UB1 plastic utility box measuring 158 × 95 × 53mm. This box is supported on a timber plinth (or base) using threaded rods and three 50mm lengths of 10mm ID aluminium tubing.

The first step is to prepare the box by drilling the various holes. We've prepared a template (in PDF format) to make this job easy. This can be downloaded from the *EPE* website and printed onto plain paper.

While you are there, download the front panel artwork and the drilling templates for the timber plinth and the smaller UB5 case. These can also be printed onto plain paper, but for a better result, print the panel artwork onto photographic paper.

Next, cut the case template sheet into its various sections, then attach the templates to the case (eg, using adhesive tape) and drill the holes to the dimensions indicated. Use a small pilot drill to start the larger holes, then carefully enlarge them to their correct sizes using larger drills and a tapered reamer.

Be careful not to over-enlarge the 10mm-diameter hole for the antenna. The aluminium antenna tube should be a tight fit into this hole.

Once the holes have all been drilled, the main label can be affixed to the lid using silicone sealant or a suitable adhesive. Allow the adhesive to dry, then cut out the various holes using a sharp craft knife. The speaker can then be secured to the inside of the lid by smearing a suitable adhesive (eg, super glue) around its outside metal frame.

Once the speaker is in place, it can be fitted with a short figure-8 connecting cable terminated in a 2-way header plug at the far end.

The main PCB is fitted into the box by first tilting it down at the front, so that the pot shafts and the switch can be slid into their respective holes. The rear of the PCB can then be pushed down into the case, after which the

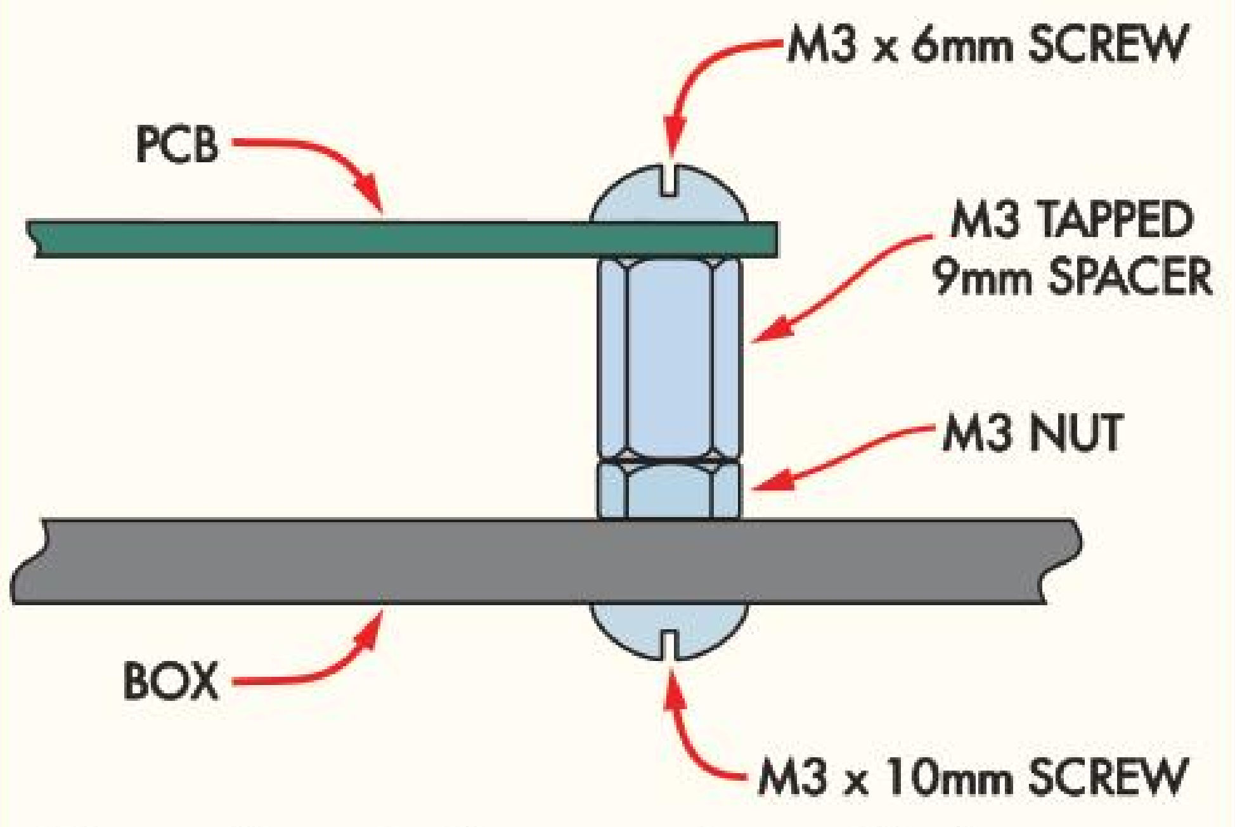


Fig.7: here's how to install the rear spacer assemblies. No spacers are required at the front of the case, since the PCB is supported along this edge by the two pot shafts.

assembly should be secured in position by attaching the nuts to the pots. Do the nuts up firmly, then mark out the positions of the two rear mounting holes on the base of the case (eg, by hand-twisting a 3mm drill through the PCB holes).

That done, remove the PCB and drill these holes in the base out to 3mm. There's no need for corresponding front mounting holes, since the PCB is supported on this side by the pot shafts.

The rear spacer assemblies can now be installed, as shown in Fig.7. First, an M3 × 6mm screw is inserted up through the bottom of the case. This is then secured with an M3 nut, after which an M3 × 9mm tapped spacer is fitted.

Don't reinstall the PCB yet – that step comes later, after attaching the case to the timber plinth.

Making the timber plinth

A piece of 151 × 90 × 19mm DAR (dressed all round) pine timber is used to make the base – see Fig.8. **Note that Fig.8 is not to scale, so you should download the full-size diagram from the *EPE* website and print it out to use as a template.**

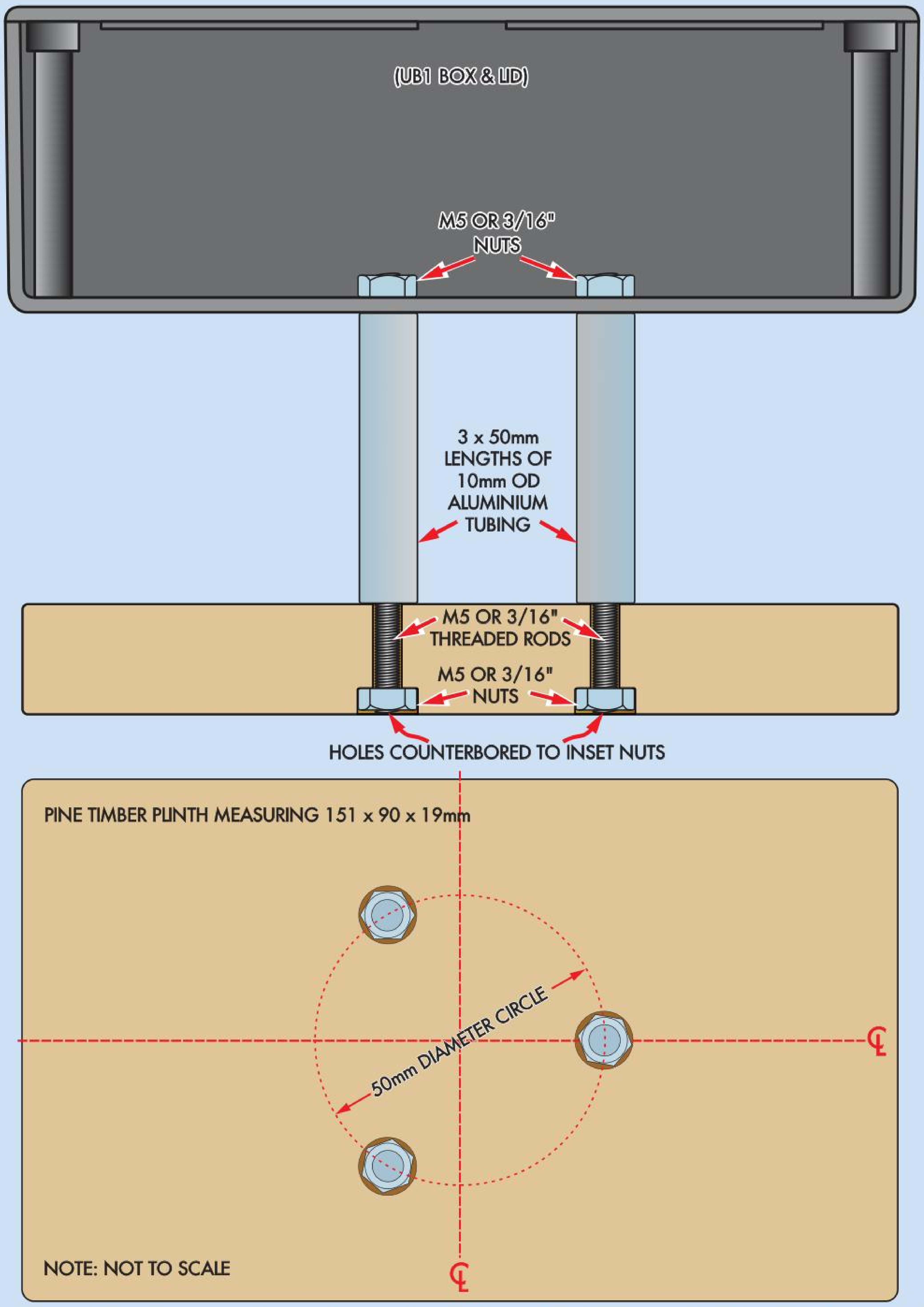

(UB1 BOX & LID)
M5 OR 3/16" NUTS
3 x 50mm LENGTHS OF 10mm OD ALUMINIUM TUBING
M5 OR 3/16" THREADED RODS
M5 OR 3/16" NUTS
HOLES COUNTERBORED TO INSET NUTS
PINE TIMBER PLINTH MEASURING 151 x 90 x 19mm
50mm DIAMETER CIRCLE
CL
CL
NOTE: NOT TO SCALE

Fig.8: this diagram shows how the timber plinth is attached to the base of the case using three 50mm lengths of 10mm-diameter aluminium tubing and M5 or 3/16-inch threaded rods. Note that this diagram is not to scale. A full-size version can be downloaded from the *EPE* website and used as a drilling template for the plinth.

Cut the timber plinth to size, then round off the edges and the corners using sandpaper. The paper template can then be attached to the base and the three holes drilled to accept either M5 or 3/16-inch threaded steel rod (zinc-plated).

Countersink the holes on the underside to allow the nuts to be recessed. Fig.8 shows the cross-sectional view (two rods only shown).

The timber base is now used as a template to mark out the corresponding holes in the bottom of the case. Drill these to suit the threaded rod, then cut the threaded steel rod into three 75mm lengths. You will also need to cut three 50mm lengths of 10mm-diameter aluminium tubing, to serve as spacers.

It's a good idea to paint the timber base black to match the box colour. After that, it's just a matter of attaching it to the case using the 75mm threaded rods, 50mm aluminium tube spacers and nuts, as shown in Fig.8.

Take care to ensure that the threaded rod protrudes no further into the box than the nut, otherwise it may later short against the tracks on the underside of the PCB.

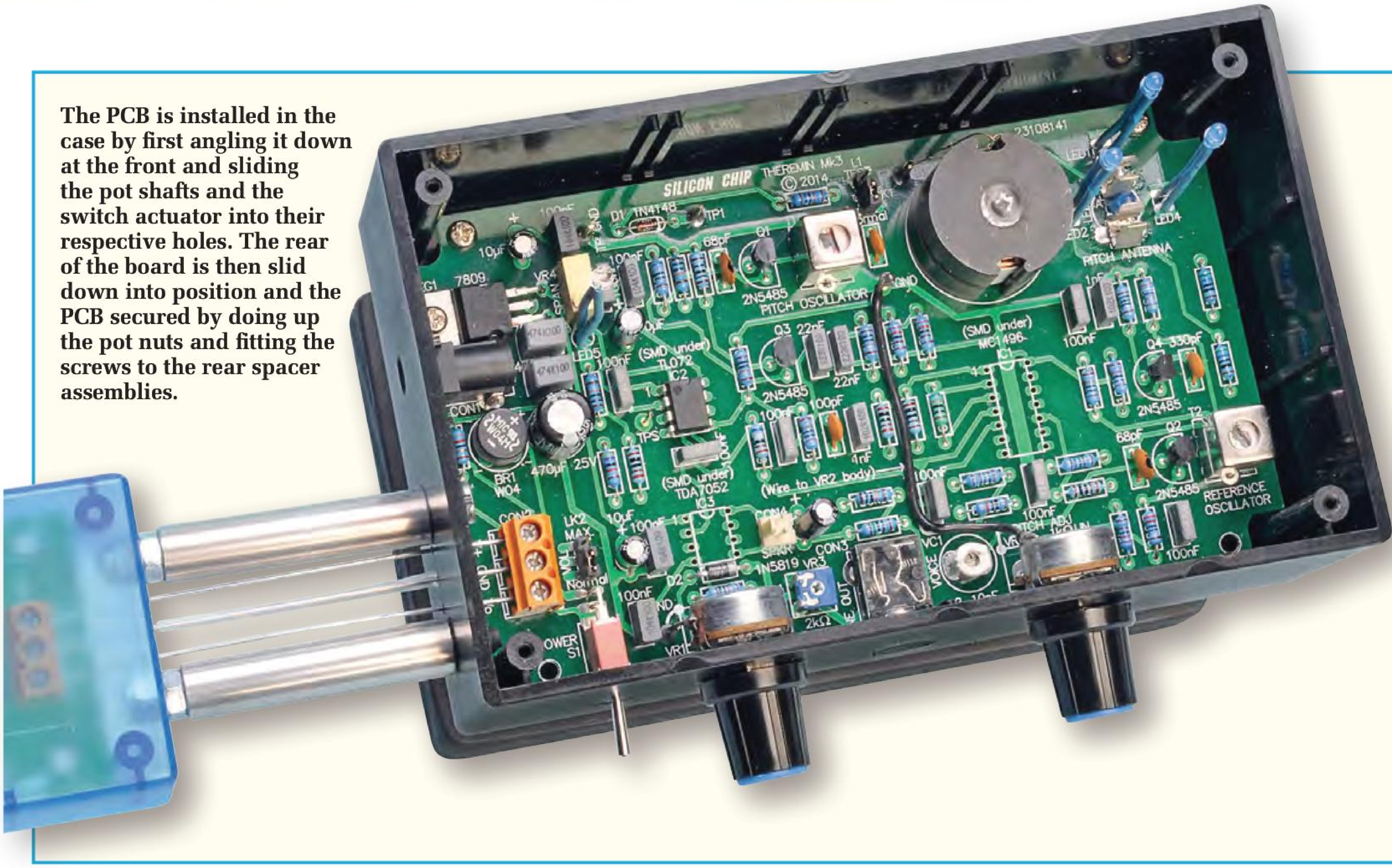
The PCB can now be reinstalled in the case and secured to the previously installed rear spacers using M3 × 6mm machine screws. Tighten these screws down firmly, then install the pot nuts and fit the two knobs. If the knob pointers are in the wrong positions, prise the end caps off and refit them so that they are correct.

Volume control case

The volume control PCB is housed in a transparent blue UB5 plastic utility box measuring 83 × 54 × 31mm. A rectangular cut-out has to be made in the base (which becomes the top) to accept the distance sensor, while five holes have to be drilled in one end for the external wiring connections and two threaded mounting rods.

As with the larger case, it's just a matter of attaching the drilling template downloaded earlier and then drilling the holes to the sizes indicated. The rectangular cut-out is made by drilling small holes around the inside perimeter, then knocking out the centre piece and filing to shape.

Be sure to make this cut-out in the base (not the lid). The case is later

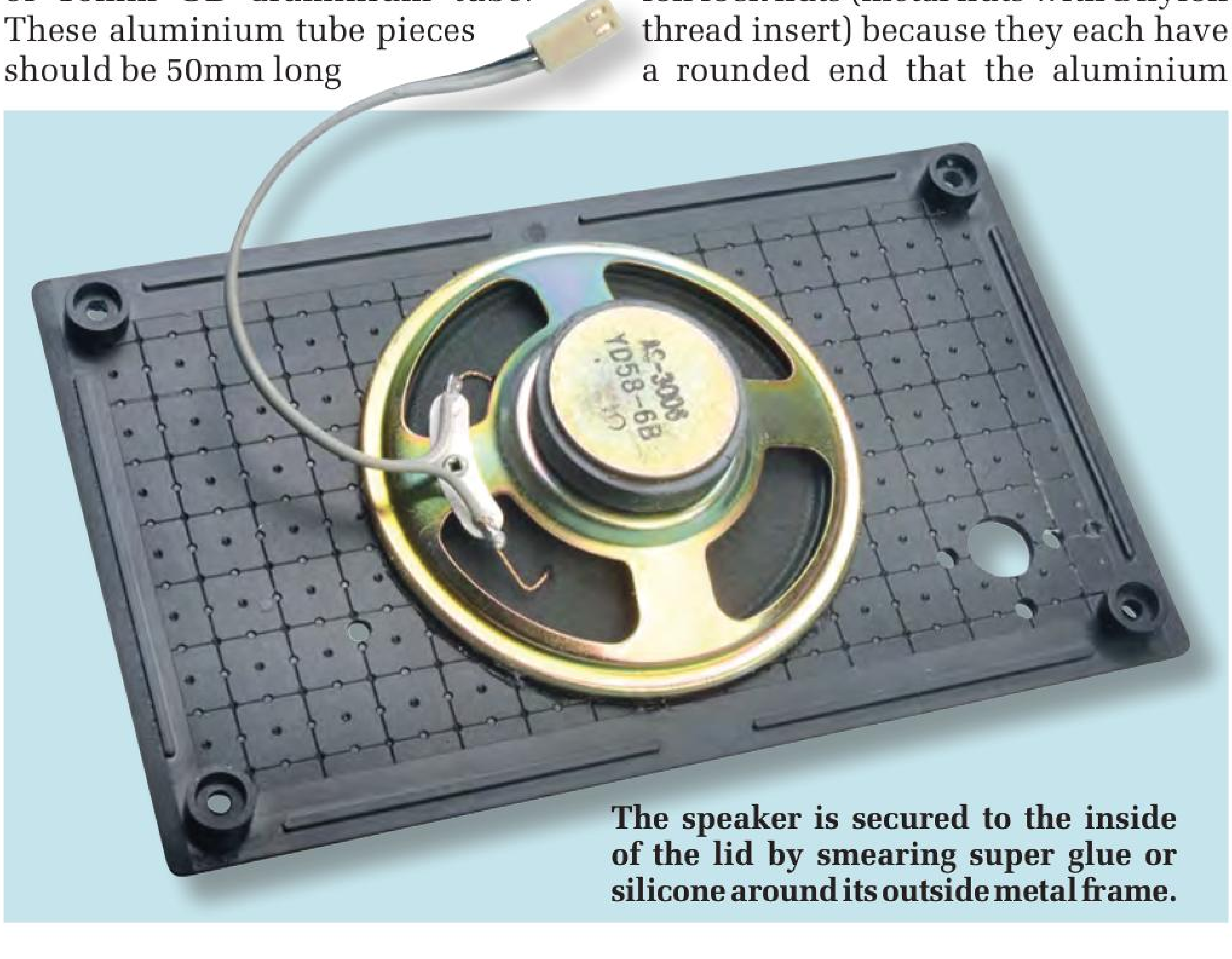


The PCB is installed in the case by first angling it down at the front and sliding the pot shafts and the switch actuator into their respective holes. The rear of the board is then slid down into position and the PCB secured by doing up the pot nuts and fitting the screws to the rear spacer assemblies.

attached to the main case with the base facing upwards and the lid on the bottom.

Refer now to Fig.9 to see how the volume control case is attached to the main case. The first job is to cut two 62mm lengths of M5 or 3/16-inch threaded rod plus two lengths of 10mm OD aluminium tube. These aluminium tube pieces should be 50mm long minus the width of the nuts used (eg, if the nuts are 4mm wide, then cut two 46mm tube lengths).

Once you have all the pieces, attach the two threaded rods to the volume control case, as shown in Fig.9; ie, for each rod, use a nut inside the case and another outside the case. We used nylon lock nuts (metal nuts with a nylon thread insert) because they each have a rounded end that the aluminium rod fits over and because they don't come undone.



The speaker is secured to the inside of the lid by smearing super glue or silicone around its outside metal frame.

The next step is to fit three straight 88mm lengths of 1mm-diameter steel wire to CON5 on the volume control PCB. These wires are then slid into the holes in the end of the case (between the nuts securing the threaded rods) and the PCB clipped into place (ie, into the integral ribs).

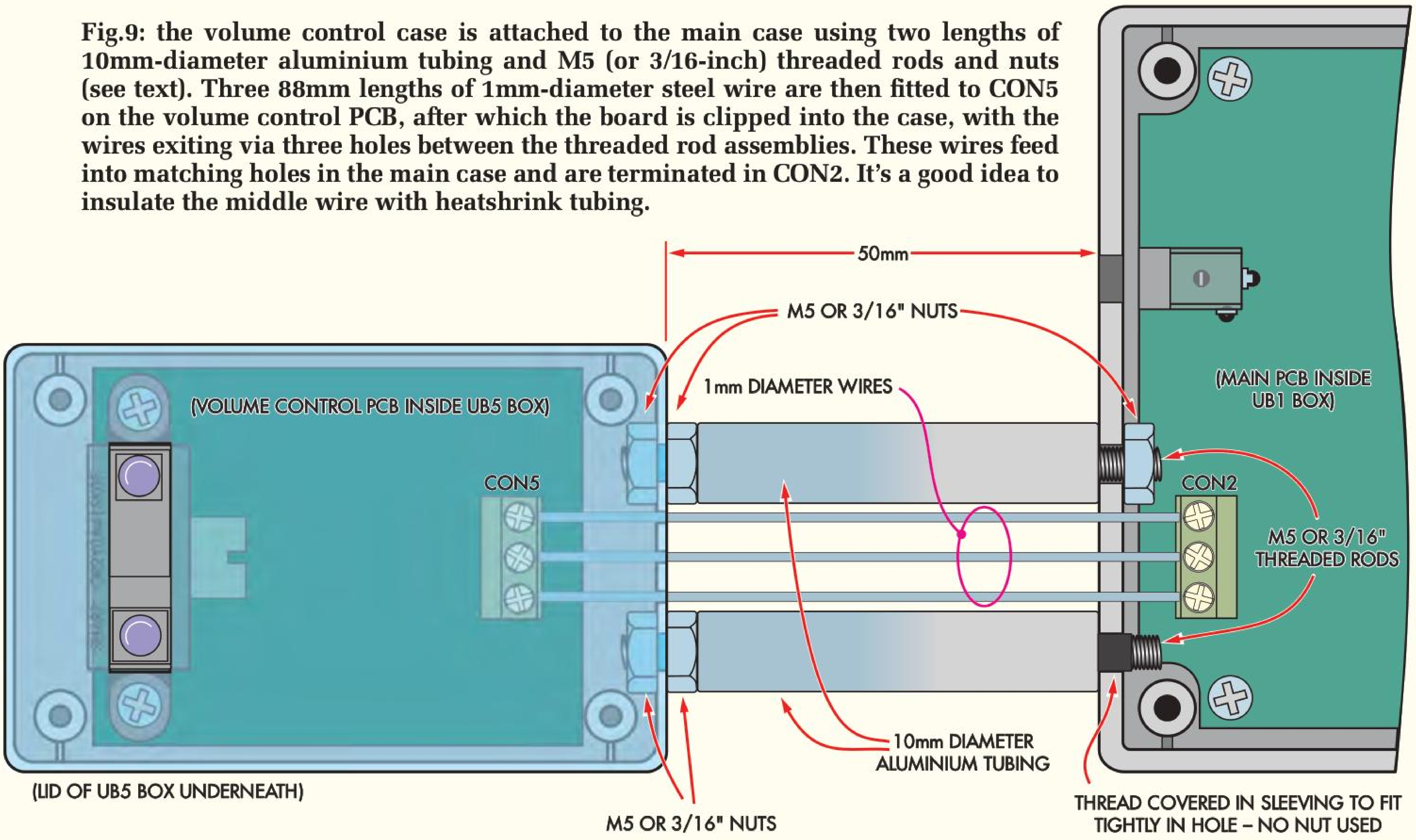
If you can't get steel wire, use 1mm-diameter tinned copper wire. This can be straightened by clamping one end in a vice and then stretching it slightly by pulling on the other end with a pair of pliers.

Final assembly

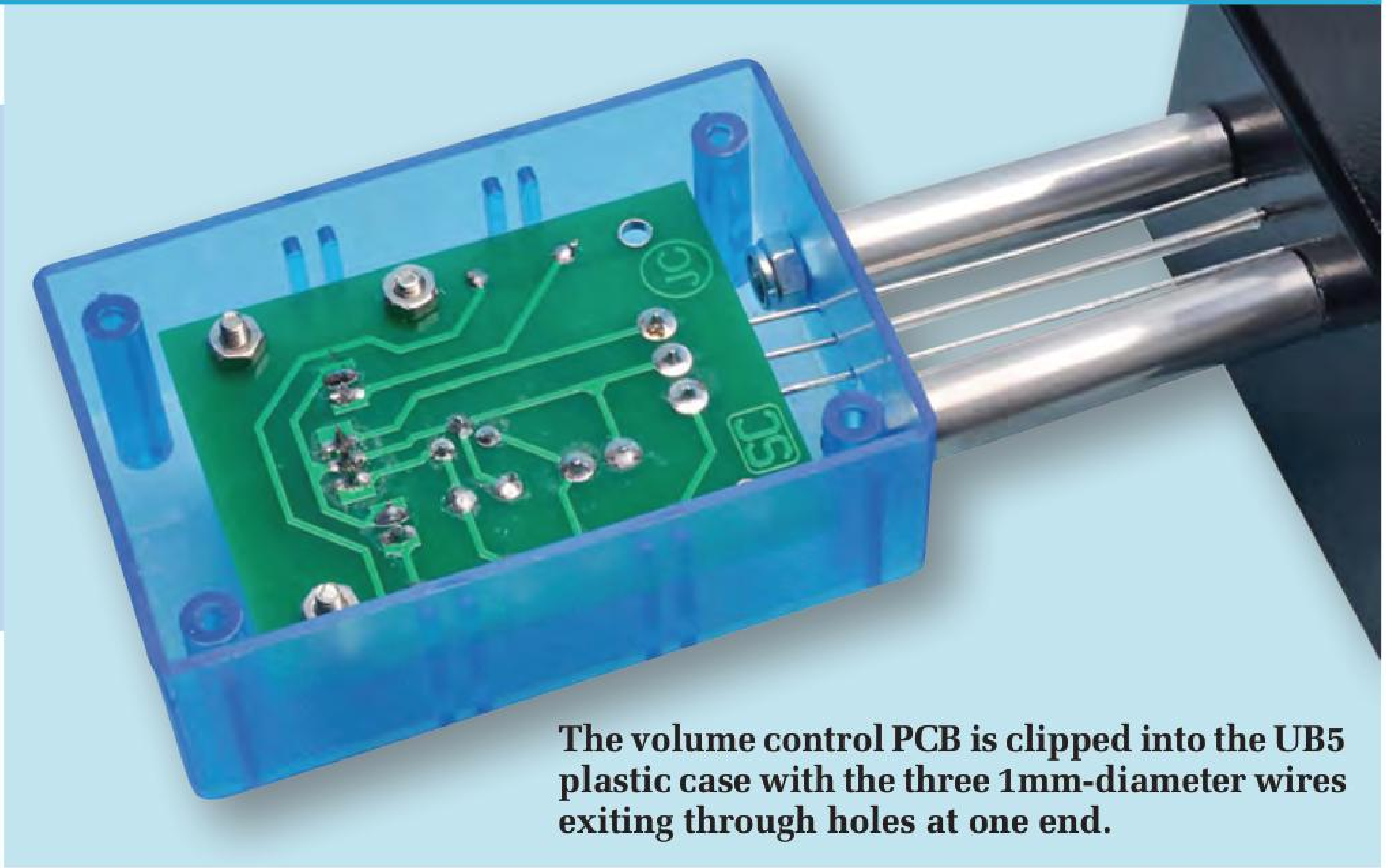
The volume control case assembly can now be attached to the main case as follows:

1) Cover the front threaded rod (ie, at the bottom in Fig.9) with a length of 6mm-diameter heatshrink tubing. This heatshrink layer should cover the entire length of the thread and can be trimmed to size after shrinking it down.
2) Cut another length of heatshrink tubing about 3mm shorter than the aluminium tubing and add this to the rod. Push it all the way up

Fig.9: the volume control case is attached to the main case using two lengths of 10mm-diameter aluminium tubing and M5 (or 3/16-inch) threaded rods and nuts (see text). Three 88mm lengths of 1mm-diameter steel wire are then fitted to CON5 on the volume control PCB, after which the board is clipped into the case, with the wires exiting via three holes between the threaded rod assemblies. These wires feed into matching holes in the main case and are terminated in CON2. It's a good idea to insulate the middle wire with heatshrink tubing.


50mm
M5 OR 3/16" NUTS
1mm DIAMETER WIRES
(VOLUME CONTROL PCB INSIDE UB5 BOX)
(MAIN PCB INSIDE UB1 BOX)
CON5
CON2
M5 OR 3/16" THREADED RODS
10mm DIAMETER ALUMINIUM TUBING
(LID OF UB5 BOX UNDERNEATH)
M5 OR 3/16" NUTS
THREAD COVERED IN SLEEVING TO FIT TIGHTLY IN HOLE – NO NUT USED

This close-up view shows how the end of one of the threaded rods is covered in heatshrink tubing so that it is a tight fit into its matching hole in the main case – see Fig.9.



The volume control PCB is clipped into the UB5 plastic case with the three 1mm-diameter wires exiting through holes at one end.

against the nut at the volume control case end before shrinking it down.

3) Repeat step 2, add more heatshrink layers until the aluminium tube is a firm fit over this threaded rod.
4) With the aluminium tubes in place, insert the three wires and the threaded rods into the main case, with the ends of the wires going into CON2. The heatshink-covered rod should be a tight fit into its hole.
5) Secure the other threaded rod with a nut on the inside of the main case.
6) Tighten CON2's screws to secure the three wires in place.

Making the pitch antenna

The pitch control antenna is also made from 10mm-diameter aluminium tubing. You also need an M4 × 10mm nylon (or polycarbonate) screw and two M4 nylon (or polycarbonate) nuts.

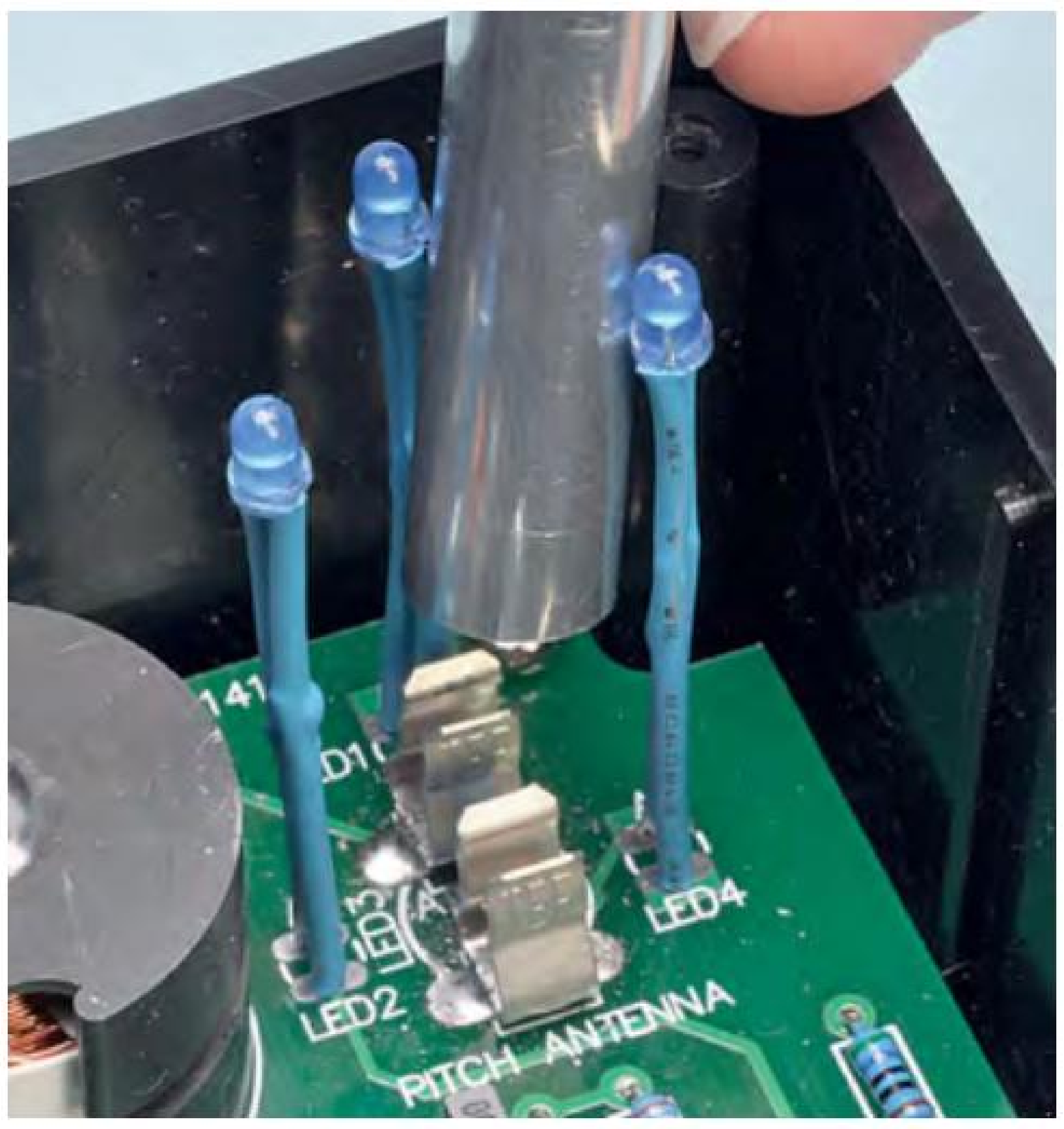
First, cut a 450mm length of the tubing and clean up the ends with a file to remove any metal burrs. That done, gently file each corner of one of the M4 nuts until it fits tightly into one end (ie, the top) of the antenna. Once it's in position, wind the second M4 nut all the way onto the screw and then screw this into the captive nut in the antenna.

This translucent 'top piece' provides the blue glow at the top of the antenna when lit by LED3 on the main PCB (ie, the blue LED that shines up the antenna tube).

The other three blue LEDs (LEDs1, 2 and 4) light the base of the antenna. As an option, these three LEDs can be covered with a translucent, hemispherical, half of a hollow ball that's slid over the antenna and pushed down onto the lid of the main case.

A ball salvaged from an empty can of roll-on deodorant is suitable. All you have to do is cut the ball in half using a fine-blade hacksaw, file the ends to

An M4 nylon (or polycarbonate) nut is pushed into the top of the pitch antenna, after which a nylon M4 screw with captive nut is fitted. This translucent assembly glows blue when lit by the LED shining up the aluminium tube.



The pitch antenna is pushed into the two fuse clips on the main PCB assembly (usually after the lid is in place). LED3 is between the two fuse clips and shines up the antenna tube to light the translucent screw assembly at the top.

a smooth finish and drill a 10mm hole in the top. It's then simply slid over the antenna.

As previously stated, the bottom end of the antenna is connected into the circuit by sliding it into the two fuse clips on the main PCB. It may be necessary to squeeze the lugs of these fuse clips together slightly so that the antenna makes a good contact. Rotating the antenna a few times will also clean the contacts if they oxidise over time.

For the time being, leave the lid off the case and simply support the antenna in its fuse clips. You are now ready for the setting-up procedure.

Setting up

The adjustment procedure is as follows:

Step 1: fit link LK1 (near the equalising coil) to the TEST position and LK2 to the MAX position.

Step 2: connect a 9VAC plugpack or a 12V DC source, switch on and check that all the LEDs light. If they don't light, check that they are oriented correctly.

Step 3: connect a DMM set to read DC volts between TPS (near IC2) and TP GND and adjust trimpot VR4 for a reading of 1.7V.

Step 4: connect the DMM between TP1 and TP GND and adjust the slug in transformer T1 for a minimum reading (note: do not use a screwdriver as this could crack the ferrite core. Either use the correct plastic alignment tool or grind down an old screwdriver so that its blade is thicker than normal and snugly fits the slot in the slug).

If you are unable to find the minimum, then either coil L1 has been incorrectly wound or its leads haven't been soldered. Check the solder joints and check also that the nylon washer spacers have been installed to provide the required 2.5mm gap between the two core halves.

Step 5: move your hand very close to the antenna (but don't touch it) and adjust T1's slug so that the voltage slightly increases. When it does, move your hand away from the antenna and check that the voltage increases even further.

If the voltage decreases instead, then the slug needs to be rotated the other way. On the prototype *Opto-Theremin*, we adjusted T1's slug for 1.1V with the hand close to the antenna and 1.7V with the hand away from the antenna.

Step 6: move jumper LK1 to the NORMAL position, connect the loudspeaker to CON4, set VR1 to mid-position and set VR2 fully clockwise.

Step 7: adjust transformer T2 until a tone is heard and set it for a low frequency. This tone should then change if you move your hand away from T2 (and away from the antenna), so this may take some trial and error.

Step 8: rotate VR2 anticlockwise and check that the pitch can be adjusted to just reach a point where there is no sound. The frequency should then become audible again and increase as a hand is brought close to the antenna. If not, reset VR2 fully clockwise again and repeat Step 7, this time adjusting T2's slug in the opposite direction.

Note that these adjustments require patience and you may need to repeat the process several times before you get it right.

Step 9: adjust VR1 fully clockwise, then adjust trimpot VR3 to limit the volume so that it isn't high enough to cause spurious vibrations or noticeable distortion.

Voicing adjustment

Trimmer capacitor VC1 must now be adjusted to set the voicing. It's just a matter of tweaking it to obtain the required sound from the Theremin.

Note that there will be a point where, at the lowest frequencies, there's a 'snap-on' effect whereby either no frequency is produced or the tone suddenly snaps on and becomes audible with hand movement. This occurs because inter-coupling between the pitch and reference oscillators causes both oscillators to track together, and if there's no frequency difference between them, there's no audible output from the mixer. However, as a hand is brought closer to the antenna, the pitch oscillator's tuning changes and it is eventually 'pulled' far enough to suddenly produce a different frequency to the reference oscillator.

Hand volume adjustment

The hand volume adjustments are all done on the main PCB as follows:

Step 1: move jumper LK2 back to the NORMAL position, then check that the volume control has a suitable hand-movement range. The volume should increase as the hand is moved away from the sensor and vice versa.

Step 2: if you want to change the range, connect a DMM between TPS and TP GND and adjust trimpot VR4 for a reading that differs from the 1.7V set earlier. Note, however, that if VR4 is set to give maximum volume too far away from the sensor, the volume will rise again at close range (ie, as the hand is brought below 40mm). This is a quirky effect of the sensor itself and is cured simply by backing off the setting for VR4.

That completes the adjustments. You can now complete the unit by attaching the lid to the main case and reinstalling the antenna, with the translucent dome slid all the way down so that it covers the three LEDs. Take care when fitting the lid to ensure that the four LEDs go through their corresponding holes. You will find it easier to do this if you apply power so that the LEDs are lit.

Finally, fit the lid to the underside of the volume control box and your *Opto-Theremin* is ready for action.

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Win a Microchip MPLAB Starter Kit for PIC24F

EVERYDAY PRACTICAL ELECTRONICS is offerring its readers the chance to win an MPLAB Starter Kit For PIC2F4. The MPLAB® Starter Kit for PIC24F (DM240015) Intelligent Integrated Analog is a comprehensive development kit featuring the PIC24F 'GC' family of 16-bit microcontrollers. This family features advanced Integrated Analog which reduces bill of materials cost, lowers noise, and has faster throughput. The board demonstrates the features of the PIC24FJ128GC010 microcontroller: direct LCD drive, 16-bit sigma-delta ADC, 12-bit pipeline ADC, 10-bit DAC, op amps, CTMU, DMA, USB, and XLP low-power consumption.

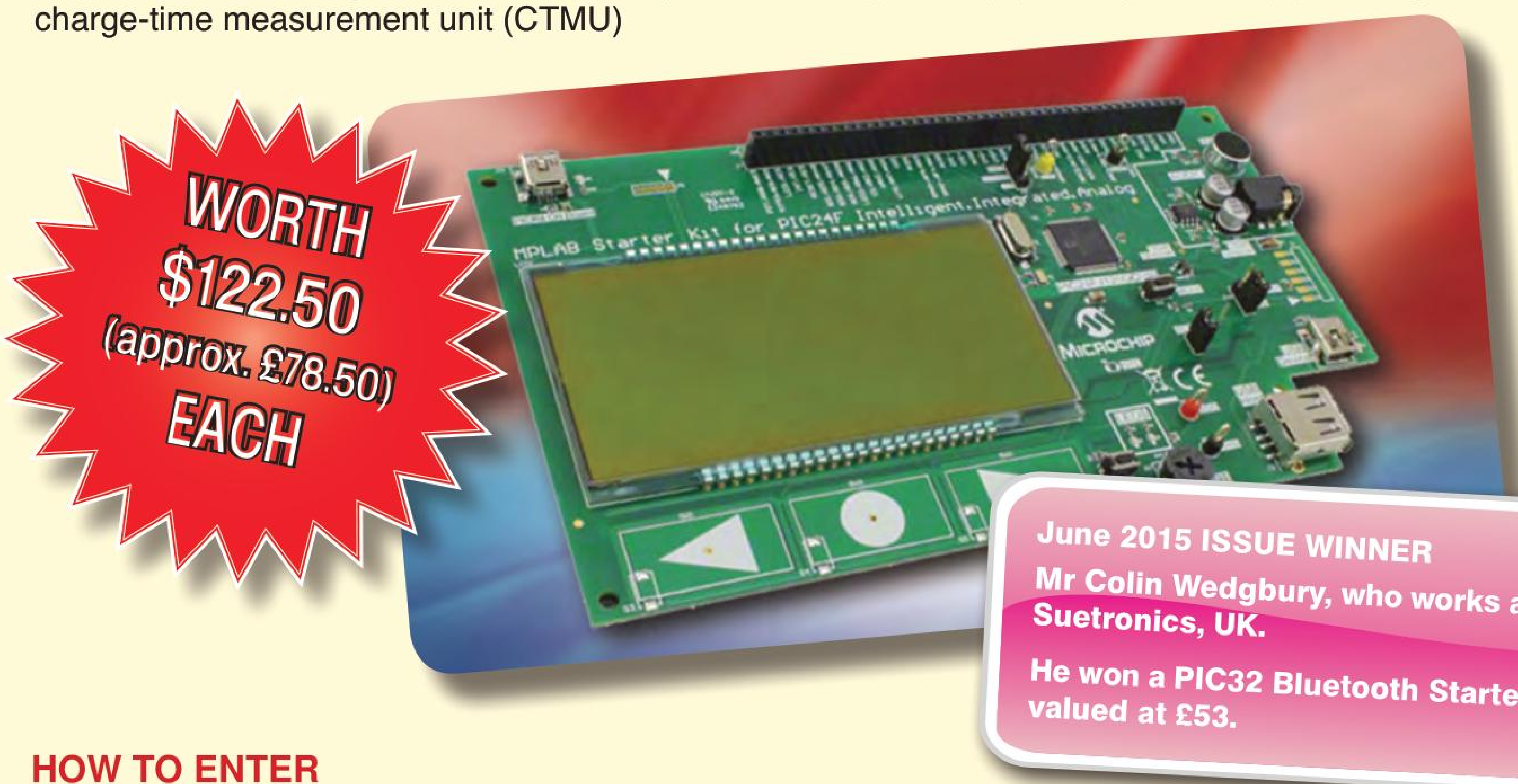
The demonstration code includes: LCD display including scrolling text with icons, sinewave audio output, light sensor, temperature sensor, resistive sensor, watch crystal-based time display, mTouch™ user input control with visual feedback, bar graph display of sensor and microphone input, and data logging of 16-bit ADC data to USB drive. The board also features an analogue connector designed to insert into a breadboard for easy access to analogue peripherals, and an expansion area for adding RF connectivity.

The board features an analogue header, allowing clean analogue signals to be accessed, preserving signal integrity. To complement the header, the board also features on-board sensors such as light sensor, potentiometer, microphone, temperature, and capacitive touch.

Additional features include USB host and device support, RF expandability, audio output via headphone jack, and on-board debugger/programmer.

KEY FEATURES

• 16-bit sigma-delta analogue to digital converter • 12-bit pipeline 10 Msps analogue-to-digital converter • 10-bit 1 Msps digital-to-analogue converter (2) • operational amplifiers (2) • comparators (3) • voltage references (3) • charge-time measurement unit (CTMU)



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Courtesy LED Lights Delay For Cars

Most modern cars have a courtesy light delay but older vehicles do not. This new circuit is specifically designed to suit LED lamps but will also work with conventional filament lamps. It keeps the interior lights of your car lit for a preset time after you shut the car doors. The lights will also turn off if the exterior lights or ignition are switched on during the time-out period.

IF YOU BUILD this courtesy light delay unit, you will be able to upgrade your vehicle to LED interior lighting. LEDs give much improved lighting compared to the yellow/orange of incandescent lamps, and the bulb diffusers will not discolour with age.

Examining older systems (designed for filament bulbs) a little analysis shows that they won't work with LED lighting unless you have at least one filament bulb connected; not the best compromise.

To understand why this is the case, look at Fig.1, which shows a typical filament bulb design concept. It's based on a MOSFET (Q1), two capacitors (C1 and C2) and a 1MΩ discharge resistor. When the door switch is closed, the interior lamp(s) light and the capacitors are discharged. The instant the door switch opens, the two capacitors charge via the filaments in the interior lighting.

Due to the different values of the two series-connected capacitors, the 47µF capacitor (C2) will charge to a voltage that's about 10-times higher than the voltage across the 470µF capacitor (C1). So with a 12V supply and taking into account the 0.7V drop across diode D1, the 47µF capacitor will have about 10.2V across it and the 470µF capacitor about 1.02V. The 10.2V across C2 becomes the gate voltage for the MOSFET, which then drives the lamps. After a short time, the gate voltage discharges via the 1MΩ resistor and the lamps go off.

As shown in Fig.1, a few refinements were often included. These include adding a short time delay to prevent MOSFET Q1 from switching on instantly when the door switch opens. This is to allow time for capacitors C1 and C2 to charge sufficiently before the MOSFET switches on and shunts the door switch. This delay is achieved using transistor Q2, which is momentarily switched on at power-up (ie, when the door switch opens) due to base drive through the 100nF capacitor and 10kΩ resistor.

When Q2 switches on, it momentarily shunts Q1's gate to ground. This prevents Q1 from conducting until the 100nF capacitor charges. The duration is only 1ms and any tendency for the lamp to briefly flicker off as the door switch opens is virtually unnoticeable.

Such designs often included additional circuitry to switch off the MOSFET (and thus turn the interior lamps off) if the tail lights were activated (ie, if the parking lights or headlights were switched on).

As stated, this circuit doesn't work with LED lighting. That's because the circuit relies on current flowing through the lamp filaments, just after they are switched off, to charge capacitors C1 and C2. Typically, a 5W lamp filament will have a resistance of about 29Ω when it is hot and so the 47µF capacitor takes much less than 1ms to charge. However, interior lighting often uses more than one lamp and so the charging resistance is usually much lower than 29Ω.

By contrast, typical 12V LED lamps incorporate two or three white LEDs connected in series with a current-limiting resistor. The voltage drop across each LED is typically 3.5V for a white LED and so the total voltage drop is around 7V with two in series, or about 10.5V with three in series. So there is not much left of the 12V supply to charge the capacitors shown in Fig.1.

When twin-LED lamps are used in this circuit, the resulting gate voltage will be around 2.9V when using a standard diode for D1 and 3.13V when

Main features

- Adjustable delay (1-133s)
- Works with both 12V LED lamps and incandescent/halogen filament lamps
- Low standby current drain
- Works with positive and negative door switch configurations
- Interior lamps go off when exterior lights or ignition are switched on

URTESY
D LIGHTS
DELAY

By JOHN CLARKE

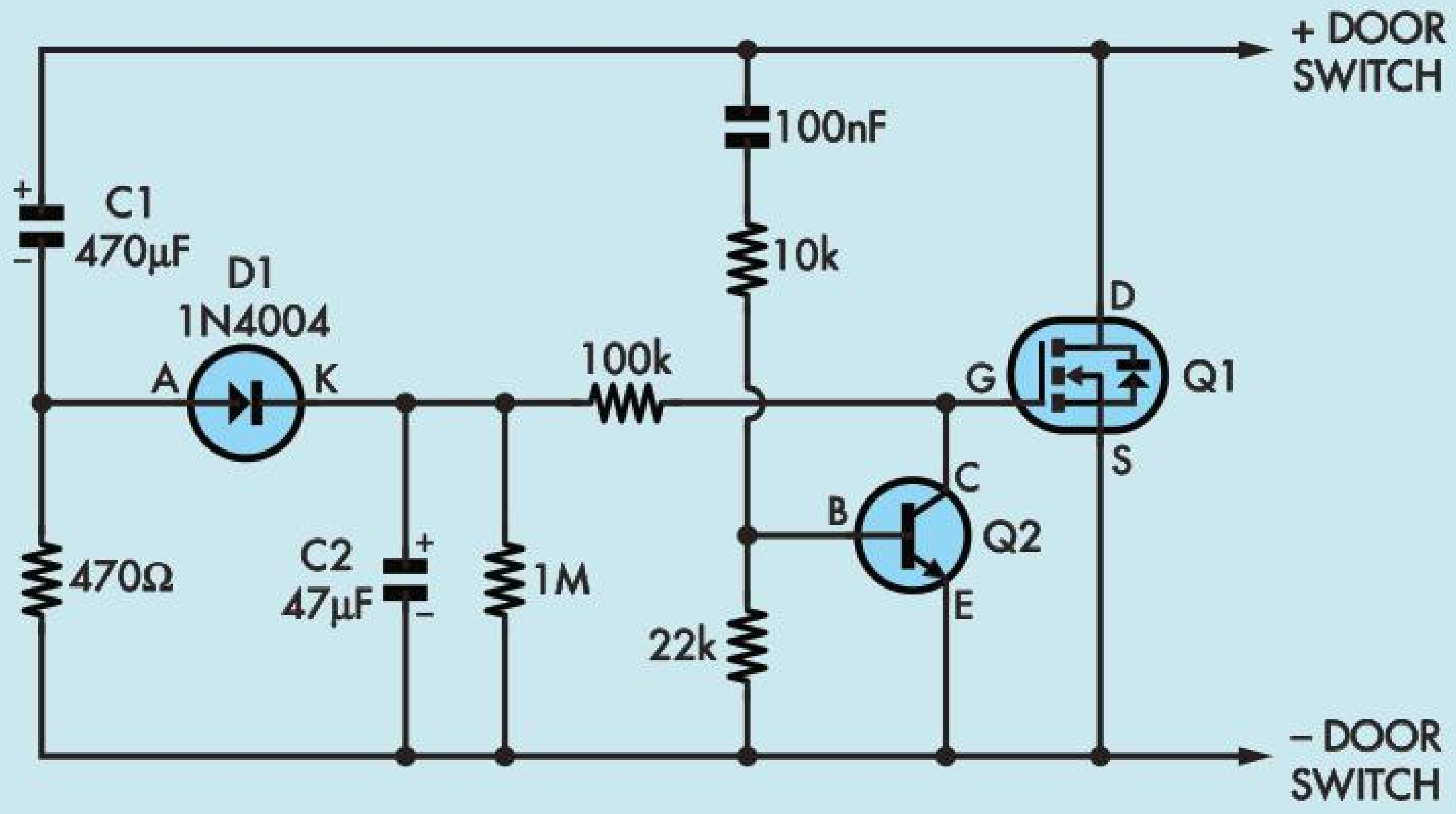


Fig.1: the basic design concept of a filament lamp courtesy light delay. It relies on current flowing through the lamp filaments to charge capacitors C1 and C2 immediately after the door switch is opened.

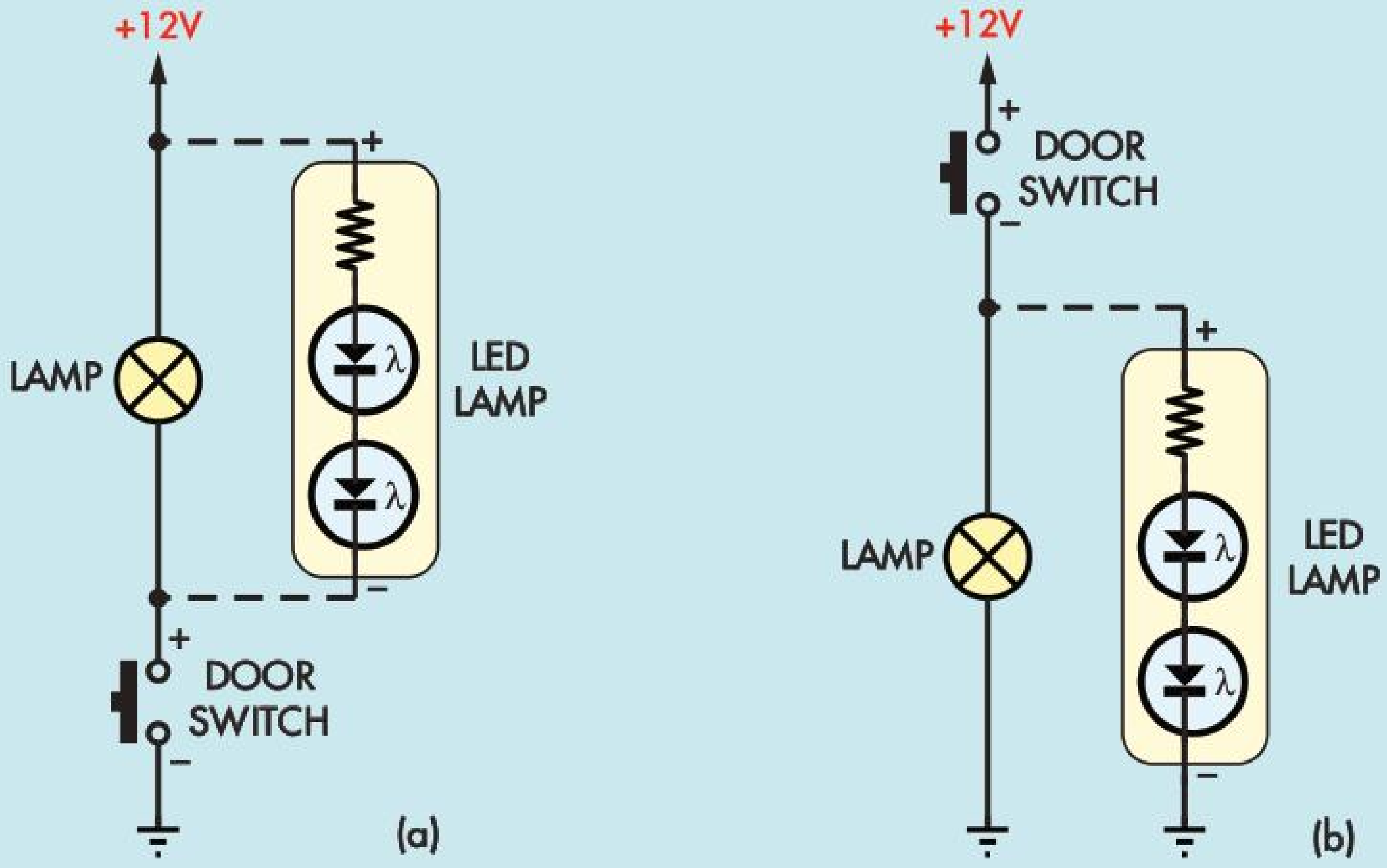


Fig.2: the door switch can either be on the chassis side of the interior lamp (a) or on the +12V side (b). Both the filament-lamp circuit and the new circuit described here work with either configuration.

Other uses

This PCB module is not just confined to vehicle use. Its circuit can also be used for timed lighting, such as in a hallway, provided you use 12V LEDs run from a 12V DC supply. A pushbutton momentary switch would be used to switch the lights on and they would then turn off automatically at the end of the preset period.

This would also be ideal for a stairwell with one or several pushbutton start switches (eg, one on each floor). With a 12V supply, up to 36W (3A) of lighting can be controlled and these could be powered from a 3A 12V power brick or similar (or use a 2A plugpack for up to 24W of lighting).

Note that the pushbutton switch needs to be rated for the total current drawn by the LED lighting.

using a Schottky diode. We do not get the expected 3.8V because of the voltage drop across the current-limiting resistor in the LED lamp.

Now, 3.13V is too low to fully switch on most MOSFETs, including typical logic-level types that can conduct (at least partially) with a 3V gate-to-source voltage but switch off below 2.5V. This means that LED lamps will not be correctly switched on by the circuit of Fig.1. Even if we substitute a MOSFET with a very low on-threshold voltage, it would be difficult to get a consistent delay period due to the low capacitor voltages compared to this threshold.

It's unfortunate that the filament-type circuit doesn't work with LED lighting because it has several desirable features. First, there's no need to connect it directly to the vehicle's 12V supply; you just connect across a door switch (in a vehicle with incandescent interior lamps) and it works. In addition, the circuit will operate regardless as to whether the door switch is on the negative side of the lamp (Fig.2a) or the positive side (Fig.2b). Provided it's connected with the correct polarity across the door switch, the circuit works in exactly the same manner for both 'high-side' and 'low-side' switching.

So how do we design a circuit to operate with LEDs? In this case, we need to connect our new circuit directly to the 12V supply as well as to a door switch. And if we want the interior lamps to switch off when the parking lights or ignition are turned on, then these too need to be monitored by the circuit.

LED version

Our *Courtesy LED Lights Delay* circuit is shown in Fig.3. Unlike the filament circuit, it also monitors the ignition as well as the exterior lights. Monitoring the parking lights or tail lights is only useful for night-time driving, since you are unlikely to use the lights during the day. By monitoring the ignition line, the courtesy lamps will immediately go out if the car is started rather than having to wait for the delay period to expire.

As with our previous circuit, the *Courtesy LED Lights Delay* operates with the door switch in either configuration (ie, high-side or low-side). Again, it's only necessary to wire the circuit to a door switch with the correct polarity. It's not necessary to know how the door switch is connected in the vehicle; you just have to identify its positive and negative leads.

The other connections to the circuit are to +12V, chassis (0V), ignition and the switched supply for the vehicle's exterior lights. The lights connection can be regarded as optional; in many cases, it will be sufficient to simply monitor the ignition line to automatically turn the interior lamps off before the delay period has ended (ie, when the car is started). The lights input connects across the parking lights or tail lights (but not the stop lights) and can be connected with either polarity.

If the courtesy lights use a low-side switching arrangement, MOSFET Q1's

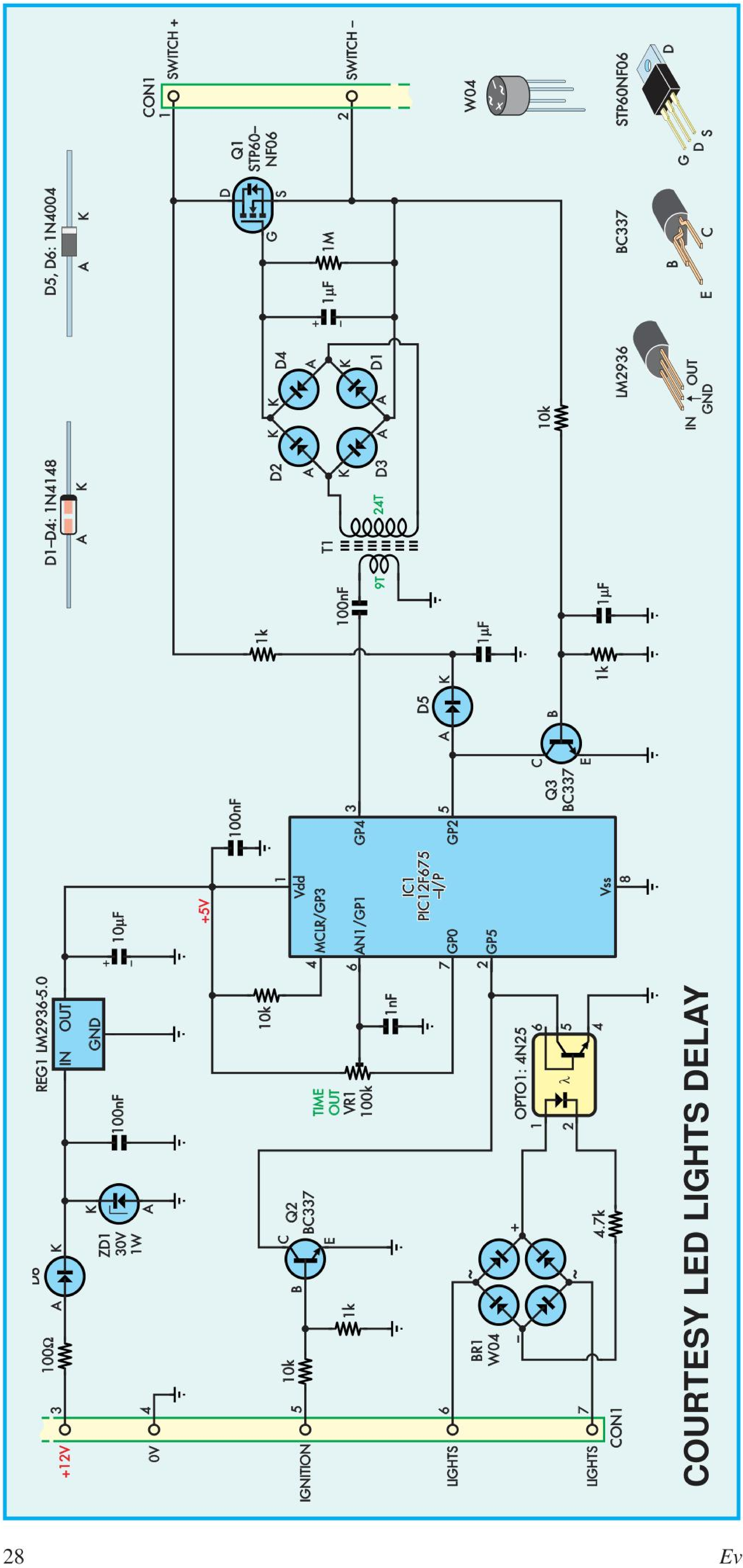

COURTESY LED LIGHTS DELAY
CON1
SWITCH +
SWITCH –
Q1 STP60–NF06
1M
1µF
D1
D2
D3
D4
T1
24T
9T
100nF
1k
10k
1µF
D5
Q3 BC337
IC1 PIC12F675 –I/P
GP4
GP2
Vdd
MCLR/GP3
AN1/GP1
GP0
GP5
Vss
+5V
10µF
REG1 LM2936-5.0
IN
OUT
GND
ZD1 30V 1W
D6
100Ω
Q2 BC337
BR1 W04
4.7k
OPTO1: 4N25
TIME OUT VR1 100k
1nF
+12V
0V
IGNITION
LIGHTS
LIGHTS
D5, D6: 1N4004
D1–D4: 1N4148
W04
STP60NF06
BC337
LM2936

Fig.3: the circuit uses PIC microcontroller IC1 to set the delay period. It also monitors the door switches, exterior lights and ignition. When a door switch is opened, IC1's GP4 output drives MOSFET Q1 (and thus the lamps) via transformer T1 and bridge rectifier D1-D4 for the delay period, as set by VR1. If the ignition or exterior lights are switched on, GP5 is pulled low (either by Q2 or OPTO1) and this immediately puts the micro to sleep and turns the interior lamps off.

source terminal will be connected to ground via SWITCH–. But this won't be the case with high-side switching.

You might expect that this could be solved by driving Q1 with an IR2125 (or similar) MOSFET driver, which could produce a suitable gate drive above the MOSFET's source voltage, whether that rises to the 12V supply (for a high-side connected MOSFET) or 0V (for a low-side connected MOSFET).

However, in the high-side configuration, this scheme relies on a low-impedance source load such as a light bulb to charge the boost capacitor during the MOSFET's off-time. This capacitor is subsequently used to generate a voltage above the 12V supply when the MOSFET switches on, so that it remains in conduction.

Once again, using LEDs for the load will mean that the capacitor will only charge to 12V minus the voltage drop of the LEDs. Ultimately, we would still be restricted to only a couple of volts for the MOSFET gate supply, so it won't work for the same reasons outlined earlier. Another problem is that the IR2125's quiescent current is rather high, at up to 1.2mA.

To get around this problem, our circuit is based on a PIC12F675-I/P microcontroller (IC1) and this drives MOSFET (Q1) via transformer T1. IC1 produces a 1MHz square-wave to drive the transformer and it provides a timing function to switch off this signal after a set period (the delay). This delay period can be adjusted using trimpot VR1.

In operation, microcontroller IC1 detects when a door switch is opened to start the delay period. It also monitors when the ignition or lights are switched on to cancel the delay period.

Because the circuit is always connected to the vehicle's 12V battery, it's vital that microcontroller IC1 has a low quiescent supply current. As a result, IC1 is normally in 'sleep' mode and draws negligible current (up to

Specifications

Delay period: adjustable from 1-133 seconds

Dim down period: 1s (can be extended – change 1μF gate capacitor on Q1)

Quiescent current: 17μA maximum, 9μA typical, 7.6μA measured (sleep mode, lamp off)

Current when running: 36mA plus LED lamp current

2µA maximum). In fact, most of the quiescent current is drawn by 3-terminal regulator REG1, as described later.

IC1's GP2 input indirectly monitors the door switch, which is wired across the MOSFET. As shown, Q1's drain connects to the positive side of the switch, while its source connects to the negative terminal.

GP2 is normally held high via an internal pull-up resistor. When the door switch is in the ground side (see Fig.2a), a closed switch pulls GP2 low via a 1kΩ resistor and diode D5. At the same time, transistor Q3 will be off since Q1's source is at ground and so Q3's base is held at 0V.

Alternatively, if the door switch is connected in the positive side of the supply, as in Fig.2b, a closed switch drives Q3's base via a 10kΩ resistor. As a result, Q3 turns on and pulls GP2 low. In this case, diode D5 is reverse biased as Q1's drain is connected to the positive supply.

So, for either connection of the door switch, IC1's GP2 input is high when the switch is open and goes low when the switch closes. IC1 is configured to generate an interrupt on a positive edge at input GP2 and when the door switch subsequently opens again, this interrupt wakes IC1 from its sleep mode. The microcontroller's firmware then starts an internal oscillator and this produces a 1MHz clock output at pin 3. This then drives transformer T1 via a 100nF capacitor.

Diodes D1-D4 rectify the voltage from T1's secondary and the resulting DC is filtered by a 1µF capacitor. This in turn switches on MOSFET Q1 to drive the interior lights, just as if a door switch was closed. Note that D1-D4 are 1N4148s because a standard bridge rectifier would not work at 1MHz.

The end result is that Q1's gate is charged sufficiently above its source to ensure it switches on, regardless of whether the source voltage is actually 0V or 12V. This configuration is known as a 'floating gate supply'.

At the same time as Q1 is switched on, IC1's GP0 output is taken low (to 0V) and this connects a 5V supply across trimpot VR1 (100kΩ). The setting at VR1's wiper is then read via IC1's AN1 input. The GP0 output is then taken high to stop the current flowing through VR1 and this is done to minimise the current drain, particularly during sleep mode.

IC1 goes to sleep again at the end of the time-out period, as set by VR1. This stops the 1MHz drive to transformer T1 and MOSFET Q1 then quickly dims the interior lights over a nominal one-second period as its 1µF gate capacitor discharges via a parallel 1MΩ resistor. Basically, the MOSFET's internal impedances rises in response to decreasing gate voltage, thereby dimming the lights until they are ultimately completely off.

Interrupting the delay

IC1 monitors the ignition and taillights circuits via its GP5 input at pin 2. If either the ignition or lights are switched on during the time-out period, the PIC immediately goes to sleep and the interior lights go out.

In greater detail, GP5 is normally held high via an internal pull-up resistor. If the ignition is switched on (eg, when the car is started), it drives the base of Q2 via a 10kΩ resistor. Q2 thus turns on and pulls GP5 (pin 2) of IC1 low to put the micro to sleep.

Alternatively, if the external lights are switched on, the resulting 12V DC supply is fed through bridge rectifier BR1 and drives the LED in optocoupler OPTO1. This in turn switches on OPTO1's output transistor, again pulling GP5 (pin 2) of IC1 low and putting the micro to sleep.

BR1 and optocoupler OPTO1 ensure that the lights circuit will work regardless of how they are switched in the vehicle. It doesn't matter whether the lights are ground connected and switched to positive or connected to positive and switched to ground.

Parts List

1 double-sided PCB, available from the *EPE PCB Service*, coded 05109141, 71 × 47mm
1 UB5 jiffy box, 83 × 54 × 31mm
1 panel-mount cable gland for 6.5mm diameter cable
1 ferrite toroid ring core, L8 material, 18mm OD, 10mm ID, 6mm high
1 8-pin DIL IC socket
1 3-way PCB-mount screw terminal block (5.08mm spacing) (CON1)
2 2-way PCB-mount screw terminal blocks (5.08mm spacing) (CON1)
1 M3 × 6mm tapped nylon spacer
1 M3 × 12mm machine screw
1 M3 nut
2 100mm cable ties
1 700mm length of 0.8mm enamelled copper wire
1 100kΩ miniature horizontal trimpot (VR1)

Semiconductors
1 PIC12F675-E/P microcontroller programmed with 0510914A.hex (IC1) (the software to programme your own PIC is available from our website www.epemag.com, alternatively pre-programmed PICs are available from www.siliconchip.com.au)
1 4N25 or 4N28 optocoupler (OPTO1)
1 LM2936-5.0 low dropout 5V regulator (REG1)
1 STP60NF06 60V N-channel Mosfet or similar (Q1)
2 BC337 NPN transistors (Q2,Q3)
1 W04 400V 1.2A bridge rectifier (BR1)
4 1N4148 diodes (D1-D4)
2 1N4004 1A diodes (D5,D6)
1 30V 1W zener diode (ZD1)

Capacitors
1 10µF 16V electrolytic
1 1µF 16V electrolytic
2 1µF monolithic multi-layer ceramic
3 100nF MKT (code 100n or 104)
1 1nF MKT (code 1n or 102)

Resistors (0.25W, 1%)
1 1MΩ
2 10kΩ
1 4.7kΩ
3 1kΩ
1 100Ω

Miscellaneous
Automotive wire, crimp connectors, quick splice connectors

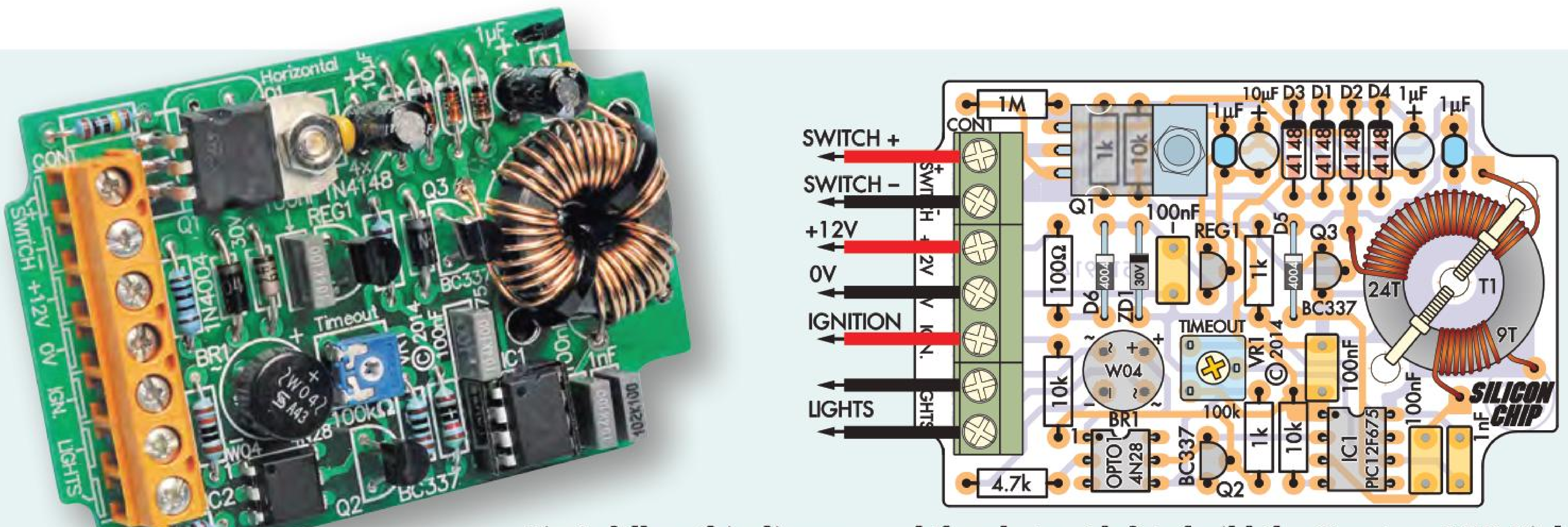


Fig.4: follow this diagram and the photo at left to build the *Courtesy LED Lights Delay*. The connection to the exterior lights circuit is optional.

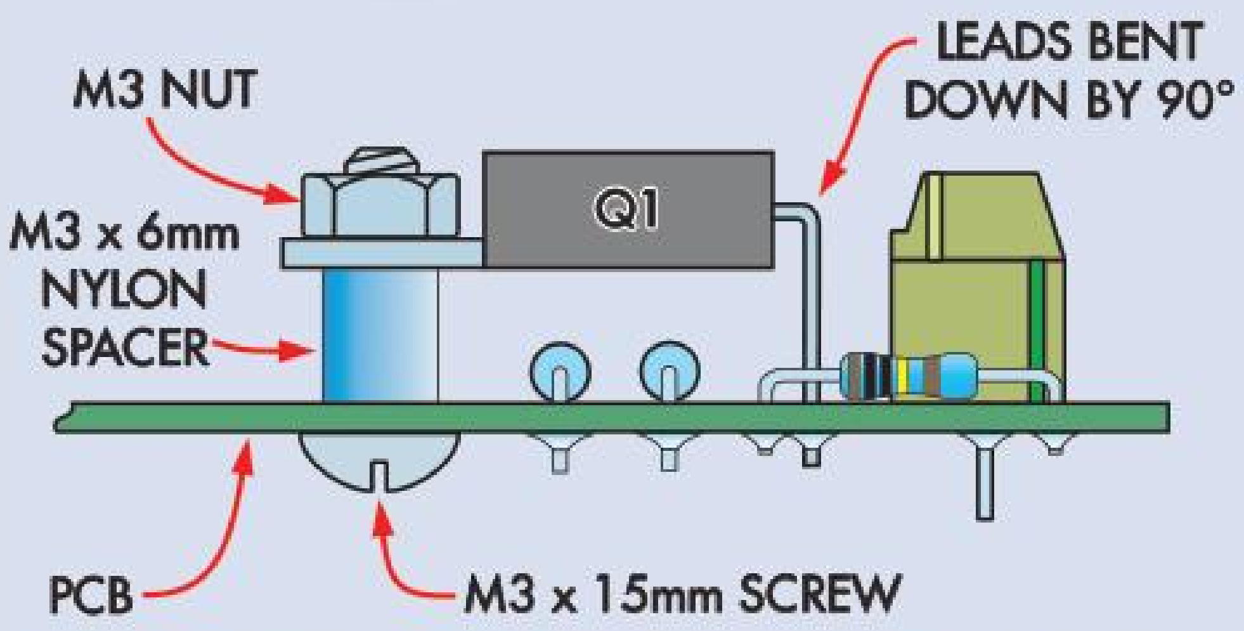
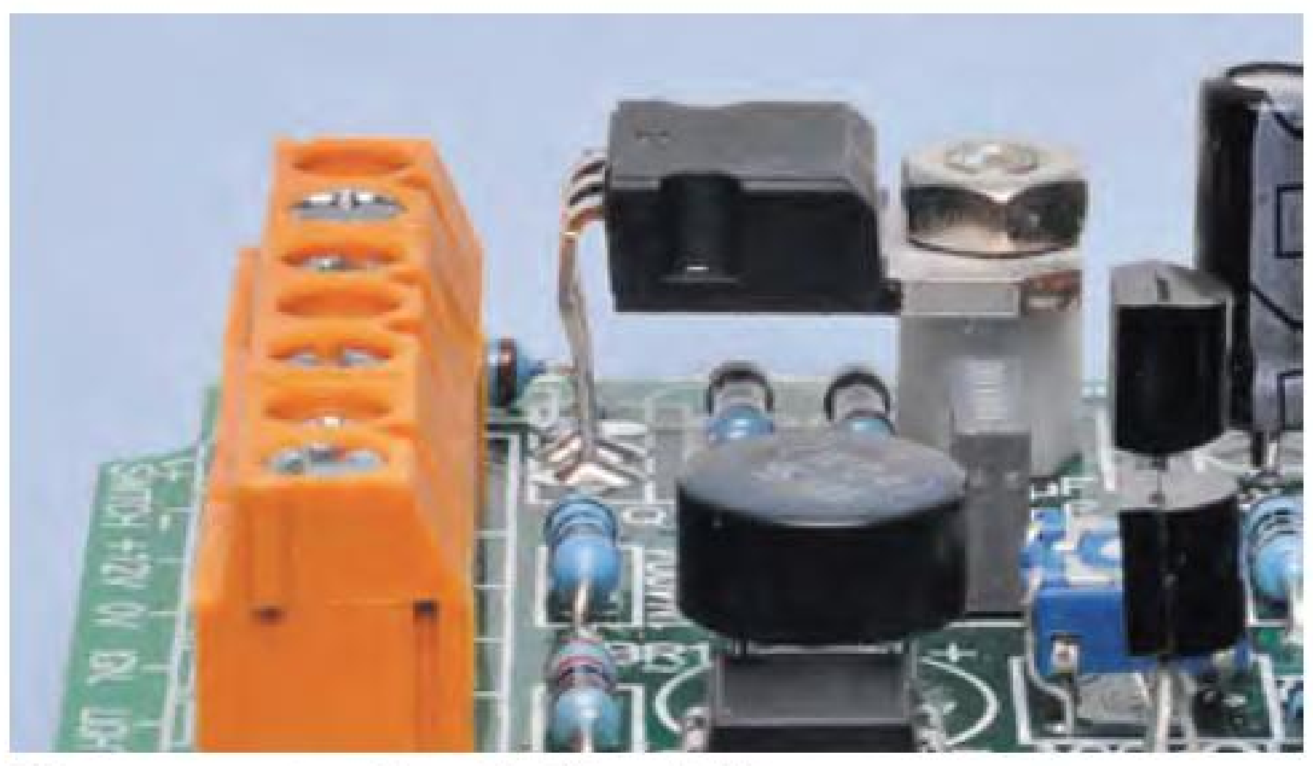


Fig.5: MOSFET Q1 is mounted horizontally, with its tab secured to an M3 x 6mm nylon spacer.



Be sure to feed the M3 × 15mm screw that secures Q1's tab to the spacer up through the bottom of the PCB.

Power supply

Power for the PIC microcontroller is derived from the vehicle's 12V supply via an LM2936 5V automotive regulator (REG1). This regulator can handle a reversed supply input and has voltage transient clamping. Diode D6, zener diode ZD1 and the 100Ω resistor are included to add extra protection. The overall quiescent current of the circuit is very low at around 9µA typical and is mainly due to the minimum current drawn by REG1.

Software

Not much is required in the way of software for IC1. As stated, it includes a rising-edge interrupt handler that wakes the PIC from sleep whenever a door switch is opened from its closed position. The PIC's internal oscillator is then automatically started and it generates the 1MHz clock signal at pin 3.

The delay counter is set from 1-133s, depending on the 8-bit ADC reading from AN1 and this period is timed using the overflow period of the internal 16-bit timer (timer 1), which occurs every 524ms.

When the delay counter reaches zero, the PIC is placed back into sleep mode so that it draws minimal power and the 1MHz clock signal ceases. During the delay period, the GP5 input is monitored and if this goes low, the processor is immediately placed in sleep mode and the LED lights quickly dim to off.

Construction

The *Courtesy LED Lights Delay* is built on a double-sided PCB, available from the *EPE PCB Service*, coded 05109141 and measuring 71 × 47mm. This clips neatly into the side channels of a UB5 plastic case and there is sufficient room to install a cable gland at the terminal block end.

Fig.4 shows the parts layout on the PCB. Install the resistors first, followed by diodes D1-D6 and zener diode ZD1. Check each resistor with a multimeter before soldering it in position and make sure that the diodes and zener diode go in with the correct polarity. The zener diode is a 30V type and will probably be marked as a 1N4751.

OPTO1, the 4N25 optocoupler, is installed next, along with an 8-pin DIL socket for IC1. Be sure to orient these parts as shown on the overlay (ie, pin 1 at top left). Transistors Q2 and Q3, regulator REG1 and bridge rectifier BR1 can now go in. Check that the LM2950-5.0 device goes in the REG1 position and check that BR1 is correctly oriented and sits flush against the PCB before soldering its leads.

The capacitors are next on the list. Watch the orientation of the electrolytics and make sure that their tops are no more than 12.5mm above the PCB, otherwise they will later foul the lid of the case. The parts list shows the codes used for the 100nF and 1nF capacitors.

Connector CON1 is made up using one 3-way and two 2-way screw terminal blocks. These should be dovetailed together to form a 7-way block which is then mounted on the PCB with the wire entry holes facing towards the adjacent edge. Make sure that this 7-way connector sits flush against the PCB before soldering the pins.

Table 1: Resistor Colour Codes

☐	No.	Value	4-Band Code (1%)	5-Band Code (1%)
☐	1	1MΩ	brown black green brown	brown black black yellow brown
☐	2	10kΩ	brown black orange brown	brown black black red brown
☐	1	4.7kΩ	yellow violet red brown	yellow violet black brown brown
☐	3	1kΩ	brown black red brown	brown black black brown brown
☐	1	100Ω	brown black brown brown	brown black black black brown

MOSFET Q1 is mounted horizontally on the PCB with its metal tab secured to an M3 × 6mm nylon spacer. To do this, first bend the MOSFET's leads down through 90° about 1mm from its body, then fit the MOSFET in position and slide the nylon spacer into position under its tab. The assembly is then secured to the PCB using an M3 × 12mm screw and nut – see Fig.5.

Note that this screw must be inserted from the underside of the PCB, so that the nut goes on top of Q1's tab. That's because the screw head is small enough not to foul adjacent PCB tracks, whereas the larger nut would run the risk of shorting out an adjacent 1µF capacitor. Do the screw up firmly to secure the assembly, *then* solder the MOSFET's leads to the PCB.

Winding T1

The PCB assembly can now be completed by winding and installing transformer T1. This transformer consists of two windings on a ferrite ring core, as shown in Fig.4. The first winding consists of nine turns of 0.8mm enamelled copper wire, while the second consists of 24 turns of 0.8mm enamelled copper wire. They are wound on opposite sections of the core and it doesn't matter in which direction they are wound.

Once the windings are in place, position the toroid on the PCB. The 9-turn winding goes through pads at the lower righthand side of the PCB, while the 24-turn winding goes to a pad just to the right of Q3 and to a pad at top right. Push the toroid all the way down onto the PCB, then secure it in position using a couple of cable ties. These pass through the centre hole of the toroid and through adjacent holes on either side.

Note that the enamel coating will need to be scraped off the wires before soldering them to the PCB.

Testing

Before installing the PIC micro, connect a 12V supply to CON1 and check that there is about 5V (4.85-5.15V) between pins 1 and 4 of IC1's socket. If this voltage is correct, disconnect the power and install IC1 with its pin 1 to the top left. If the voltage is incorrect, check the orientation of D6, the value of the series 100Ω resistor and that REG1 is an LM2936-5.0.

If you have a spare 12V LED lamp, this can be used to test the circuit before installing it in the vehicle. **Do not use a white LED on its own. It must be a LED lamp with a limiting resistor to keep the current to a safe level for the LEDs.**

Assuming you have a spare LED lamp, connect it between the switch minus terminal (pin 2 of CON1) and 0V (pin 4 of CON1). **Note that the polarity is important here – the anode or positive side of the LED lamp must go to the switch minus terminal.**

Now reapply power – the LED lamp should light for a second or so, then quickly dim to off. If that checks out, momentarily bridge the switch terminals on CON1 (pins 1 and 2). The LED lamp should now light for the length of time set by trimpot VR1 (note: timing begins when the door switch opens; ie, when the door is closed).

Assuming it works as expected, VR1 can now be adjusted to set the required delay. This ranges from 1-133s but note that the circuit's response to the trimpot setting is non-linear. The fully anticlockwise to mid-position setting has the range of 1-33s, while the next half of the travel is divided into two equal sections. The first section has a range from 33-66s, while the remaining clockwise section sets the delay from 66-133s.

Having set the delay period, you can test that the ignition input works. That's done by first triggering the delay period, then connecting the ignition input (pin 5 of CON1) to +12V using a wire link. When you do this, the lamp should extinguish after one second or so.

Note that you will need to set a reasonably long delay time for this test, to give yourself time to connect the ignition input to the +12V terminal.

Similarly, you can check that the lights inputs work by triggering the delay and connecting either pin 6 or pin 7 of CON1 to +12V and the other pin to ground. Again, the LED lamp should turn off after a second or so.

If you wish, you can increase this 1s dim-down period by increasing the 1µF electrolytic capacitor value at the Q1 gate. It will be around 10s with a 10µF capacitor. Note that this dimming period is additional to the time-out or delay period. So if the time-out is set at 15s, the overall LED 'on-period' will be 15s plus the dimming period. For 1s dimming, the total time-out will be 16s.

Final assembly

If it all works as expected, drill a 12.5mm hole in the end of the UB5 box for the cable gland. This hole should be positioned 13mm down from the top of the box and centred horizontally. Drill a small pilot hole to begin with, then carefully enlarge it to size using a tapered reamer and mount the cable gland in position.

The assembled PCB is now simply clipped into the UB5 box with CON1 adjacent to the cable gland. This gland clamps the external wiring cable to prevent the connecting wires from being pulled out of CON1.

Installation

To connect the unit, you will need to access one of the door switches, +12V power, the ignition line and either the tail light or parking light connections. Alternatively, you may wish to just use the ignition input and not bother with the lights input.

Note that some door switches will have two wires while others have only a single wire connection. In the latter case, one contact is connected directly to chassis at the switch mounting position.

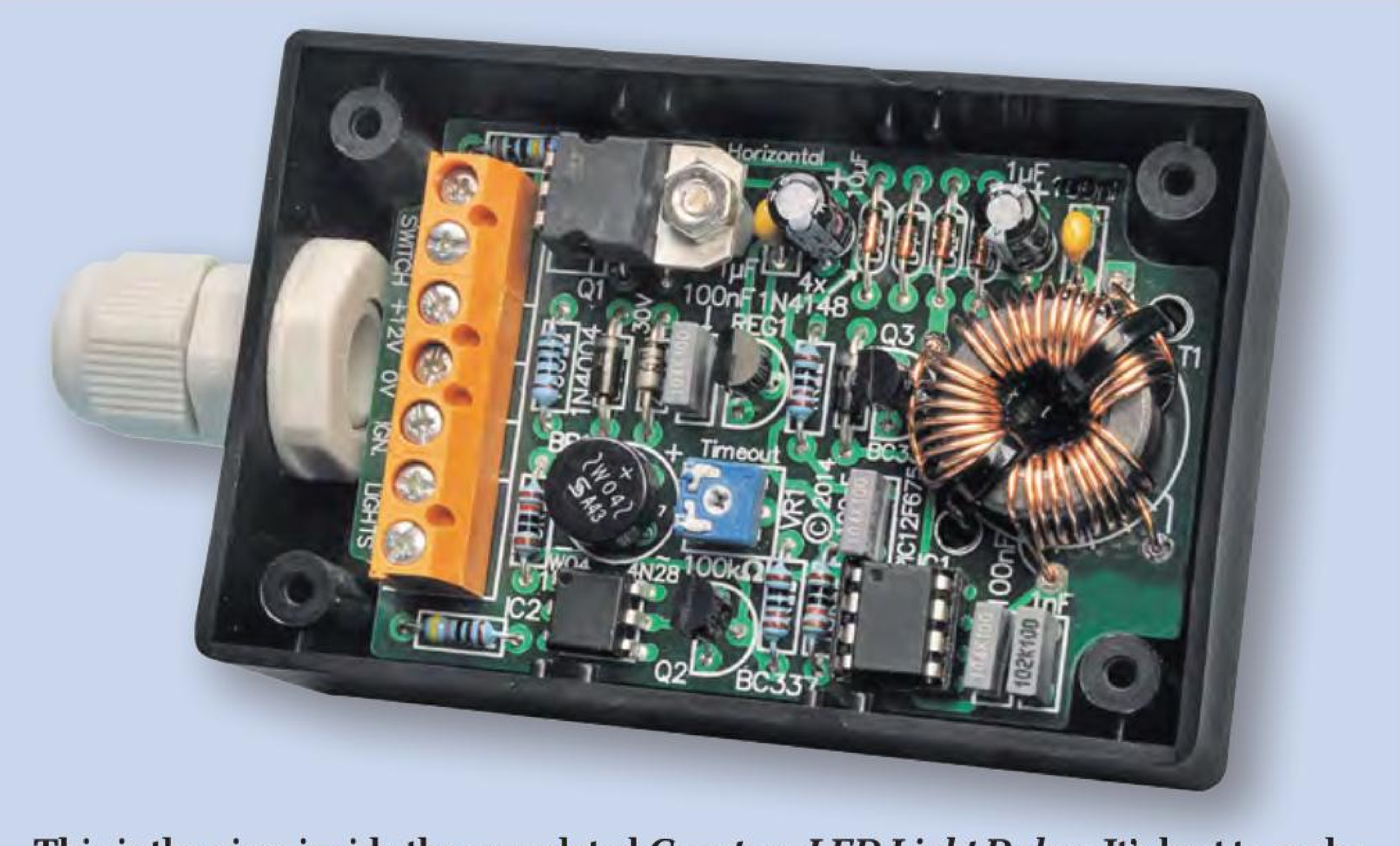
It's important to get the door switch connections to the unit the right way around. The positive door switch connection must go to the switch positive of the *Courtesy LED Light Delay*. You can quickly determine which is the positive door switch connection by using a multimeter to measure the voltage across the door switch when it is pushed open.

Note that if there's only a single wire running to the switch, this will be the positive (assuming the chassis connection is negative).

For the +12V supply rail, you will need to find a source of +12V that remains on when the ignition is off. This +12V supply rail must be protected by a fuse in the vehicle's fusebox and is best derived at the fusebox itself. The 0V lead can be run to an eyelet connector that's screwed to the chassis.

The lights terminals on the *Courtesy LED Lights Delay* are connected across one of the tail lights or parking lights. You can access this wiring either directly at the lights socket wiring, at the lights switch or in the fusebox. It doesn't matter which way around you connect them, since the bridge rectifier automatically caters for both polarities.

Once you have found the relevant wiring points, it's a good idea to disconnect the vehicle's battery before running the wiring, to guard against



This is the view inside the completed *Courtesy LED Light Delay*. It's best to make the external wiring connections to CON1 with the PCB out of the case. The wires are then pushed back out through the cable clamp as the PCB is clipped into position.

any inadvertent short circuits. Note that all wiring should be run using proper automotive cable and connectors.

Once the wiring is complete, reconnect the battery and check that the courtesy lights remain on after the door is closed. Now turn the ignition (or the exterior lights) on and the courtesy lights should quickly dim to off (over 1s or so).

Finally, the unit can be mounted in any convenient location under the dashboard. It's up to you how you secure it, since a suitable position will vary from vehicle to vehicle.

Existing delay circuit

What if your vehicle already has a courtesy lights delay? This may work fine if you substitute LED interior lamps for your car's original incandescent lamps, but there's always a possibility that it may not. In that case, you may wish to use the *Courtesy LED Lights Delay* instead.

One problem here is that the door switches will probably be connected to the existing delay circuit rather than directly to the interior lamps. Bypassing this delay circuit will therefore involve disconnecting all the door switches and wiring them directly to the interior lamps instead.

That's too complicated (and time consuming) to be practical in most cases, but there is a way around this – keep the original delay circuit and simply add the *Courtesy LED Lights Delay* unit to the existing installation. That's done by connecting the delay unit in parallel with the existing unit across one of the door switches.

There's just one wrinkle to watch out for here – the original delay circuit may pull one side of the door switch to +5V rather than +12V. This should be checked using a multimeter and if it does go to +5V, the 1kΩ pull-down resistor connected to Q3's base will have to be increased to 10kΩ (otherwise the transistor won't turn on).

Note that, depending on the circuit used, the original delay period may be added to the delay introduced by the *Courtesy LED Lights Delay* unit. That won't be a problem, however, since the *Courtesy LED Lights Delay* period can be adjusted down to as low as 1s.

Note also that connecting the *Courtesy LED Lights Delay* in parallel with an existing delay circuit may not work in all cases. It will very much depend on the vehicle and the circuit used.

Troubleshooting

If the courtesy lights are always on, the door switch terminals have probably been connected to CON1 (at pins 1 and 2) with reverse polarity. If that happens, the courtesy lights turn on via the intrinsic reverse diode inside MOSFET Q1 and simply swapping the leads to the door switch will fix the problem.

Finally, if the interior lights switch off immediately after the door is closed (and the connections are correct), check that there is no voltage applied to either the lights terminals or the ignition terminal on CON1 (pin 5, 6 and 7).

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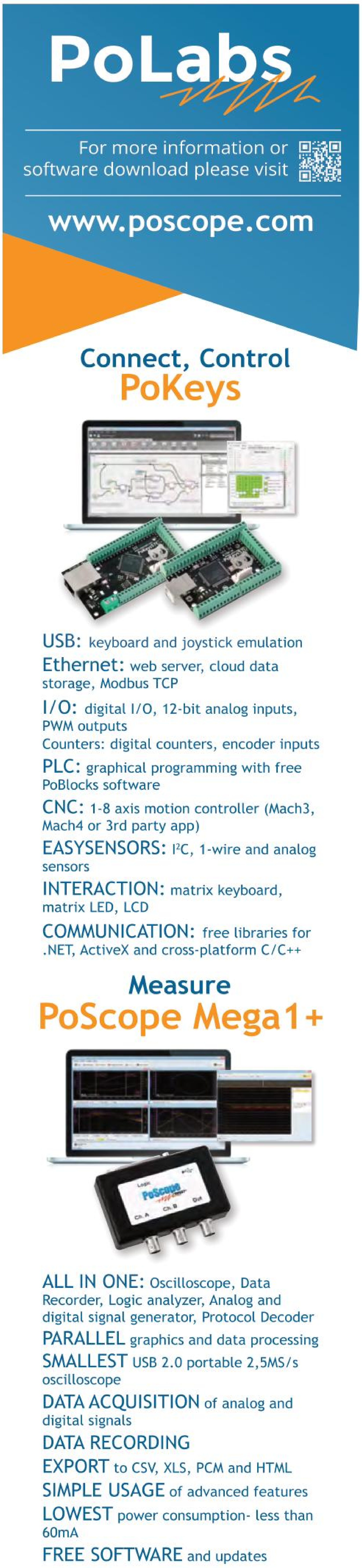
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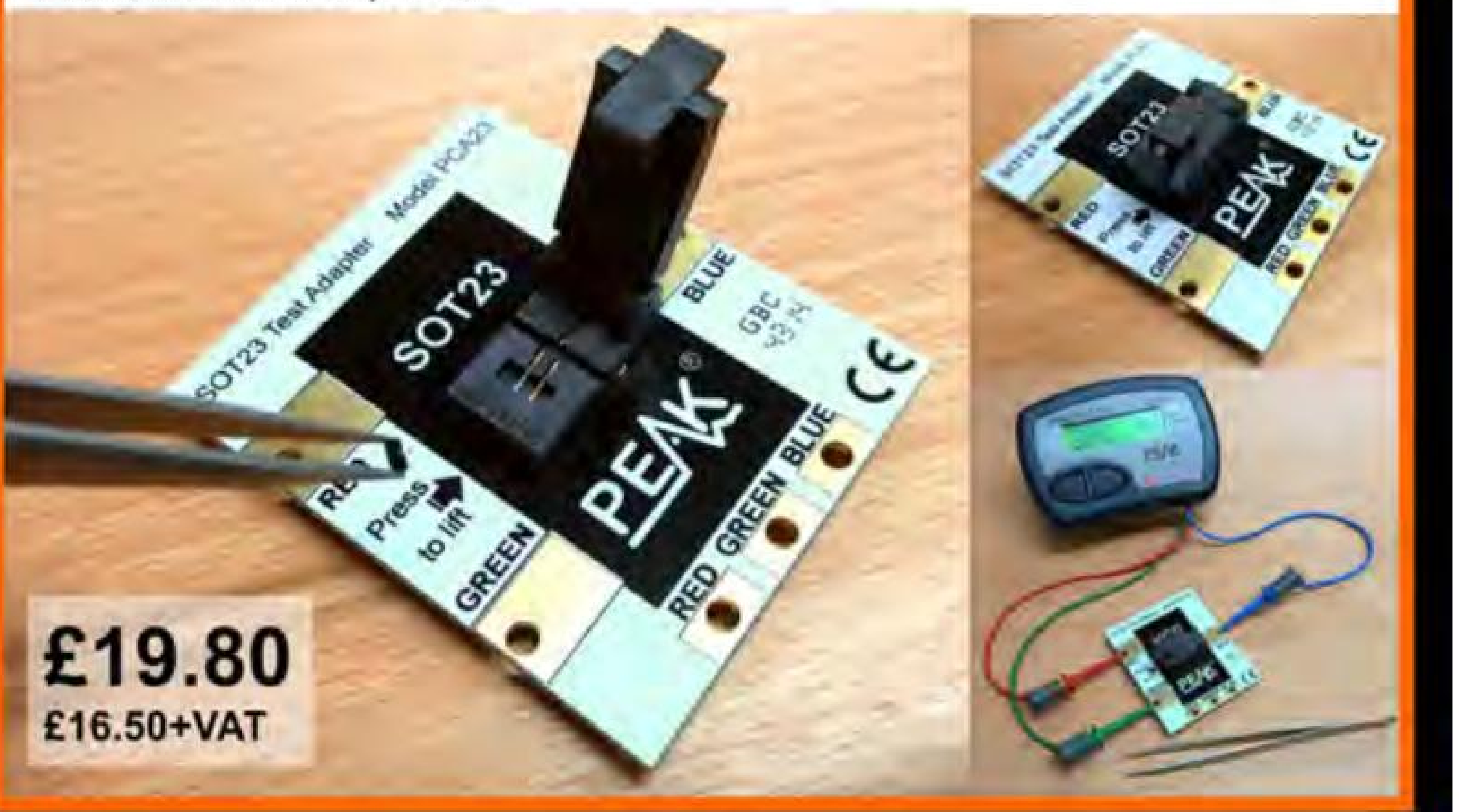

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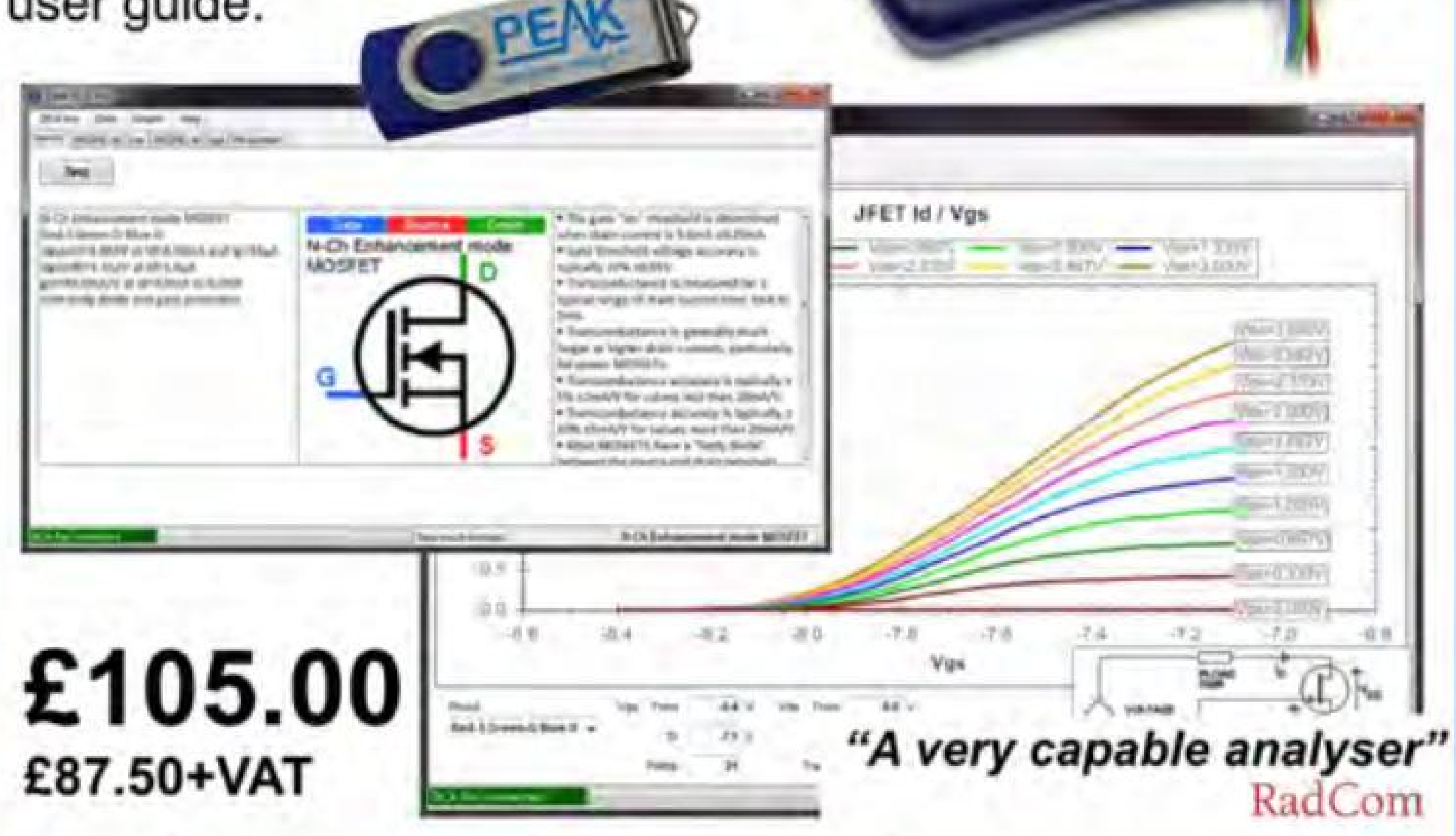

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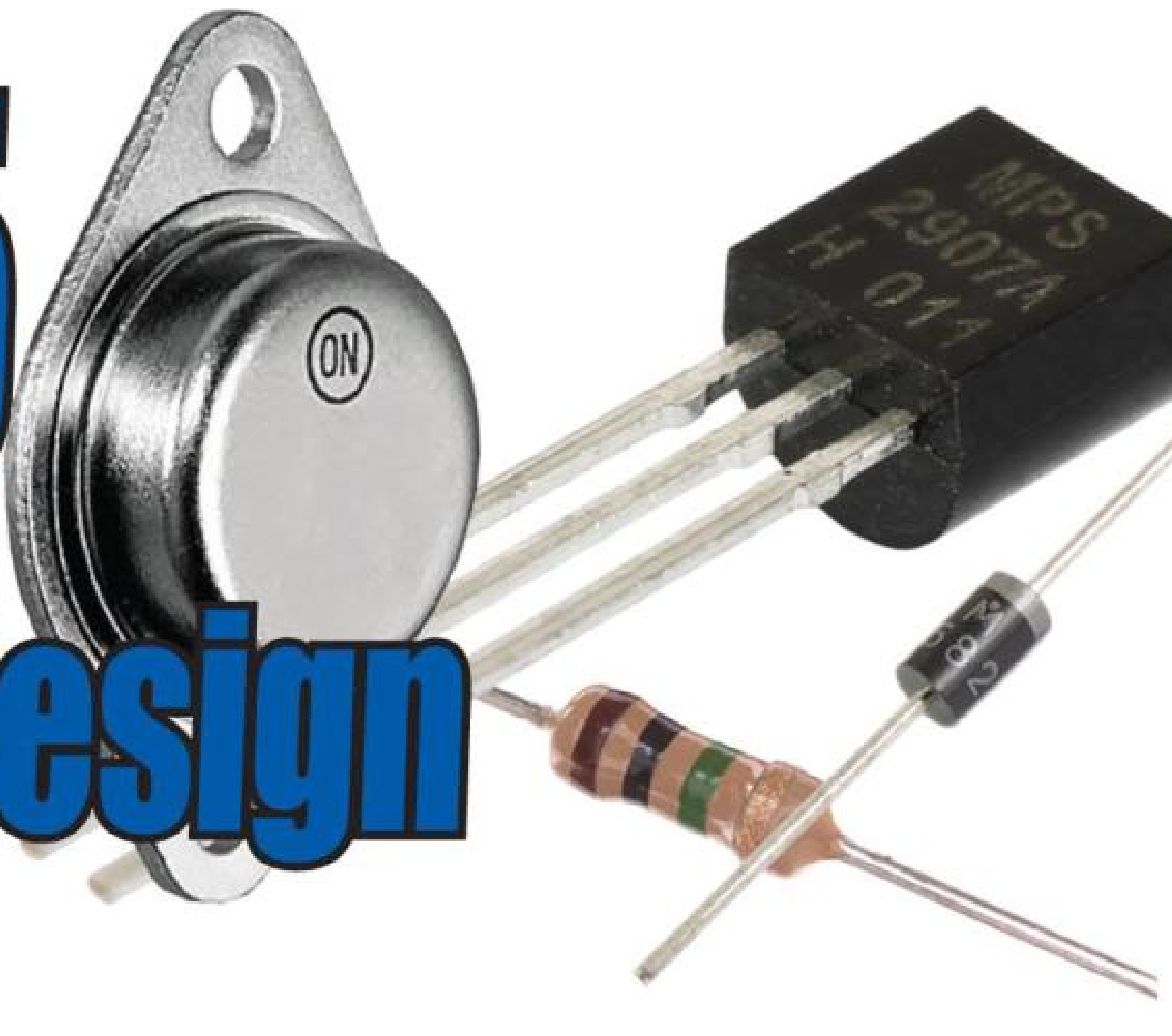
Teach-In 2015

Discrete Linear Circuit Design

Part 9: Bringing it all together



by Mike and Richard Tooley

Welcome to *Teach-In 2015*. This series is aimed at anyone wishing to develop a detailed understanding of linear discrete semiconductor devices and how they are used in a diverse range of circuits. We hope you will join us on this exciting voyage of discovery! Each part of our series is devoted to a different aspect of discrete linear circuit design such as modelling and simulation, measurement and testing, noise and distortion. Last month, *Knowledge Base* introduced negative feedback and explained how it provides an effective way of making an amplifier both stable and predictable. *Discover* was devoted to output stage design and it introduced several important circuit building blocks, including Darlington output stages, quasi-complementary configurations and compound feedback Sziklai pairs.

Introduction

In this, the penultimate part of *Teach-In 2015*, we shall be bringing together a number of ideas and concepts from previous months. *Knowledge Base* takes a detailed look at stability, thermal and over-current protection, while *Get Real* is devoted to the design and construction of a low-cost, high-quality 10W power amplifier that will out-perform most of today's similarly rated integrated circuit amplifiers. This final practical project is designed for use with previous *Get Real* circuits and it will help you to build your own good-quality modular audio system.

Knowledge base: Improving amplifier performance

Loudspeakers as loads

Earlier in this series we considered the load present at the output of an amplifier, and the power that would appear in it. However, when we did this we assumed that the load was well-behaved and just acted like a pure resistor and had no reactive components. Unfortunately, in the real world this is very far from the truth since most loudspeakers exhibit an impedance that varies widely over the audio frequency range. To illustrate this, we carried out some measurements on a popular Sharp XL-60H 40W loudspeaker system (see Fig.9.1). The impedance/frequency characteristic of this relatively low-cost system is shown in Fig.9.2a.

There are a few important things to note from Fig.9.2a. First, the distinct and relatively sharp peak in impedance at around 150Hz and second, the steady increase in impedance that occurs above 4kHz. That said, the system manages to maintain its nominal 4Ω impedance from 20Hz to 120Hz and from 250Hz to 4kHz. These variations of impedance are 'seen' by the output of the amplifier, but they are some way from the nominal 4Ω that might at first be assumed.

So, how do these undesirable fluctuations arise and what can be done to minimise them so? To partly answer this question, we carried out some measurements on the Sharp XL-60H loudspeaker using an accurate universal LCR bridge. Our measurements revealed that the loudspeaker had a DC resistance of almost exactly 4Ω, together with a series inductance of 130μH. This latter value was in very close agreement with 129μH calculated from our previous impedance measurements.

Resonance

The low-frequency peak (around 180Hz) in impedance results from resonance in the low-frequency driver unit. The size of the resonant peak and the frequency at which it occurs is determined not only by the characteristics of the driver (mass, suspension) but also on the enclosure in which the driver is placed. The free-air resonance is often quoted in a driver's specification and the size of the resonant peak relates to the Q-factor of the unit. The larger the Q-factor, the sharper the peak will be. When a driver is placed in an enclosure, its resonant frequency will increase along with its Q-factor. A typical increase in resonant frequency might be around 30Hz for a small sealed enclosure fitted with a 6-inch driver. At the same time, the Q-factor might increase from around 1.2 to a little more than 1.5. It should go without saying that loudspeaker design is a complex art, and in most cases it can be instrumental in making significant improvements to the quality of the sound produced by an audio system.

Voice coil reactance

The steady increase in impedance above 4kHz is attributable to the inductive

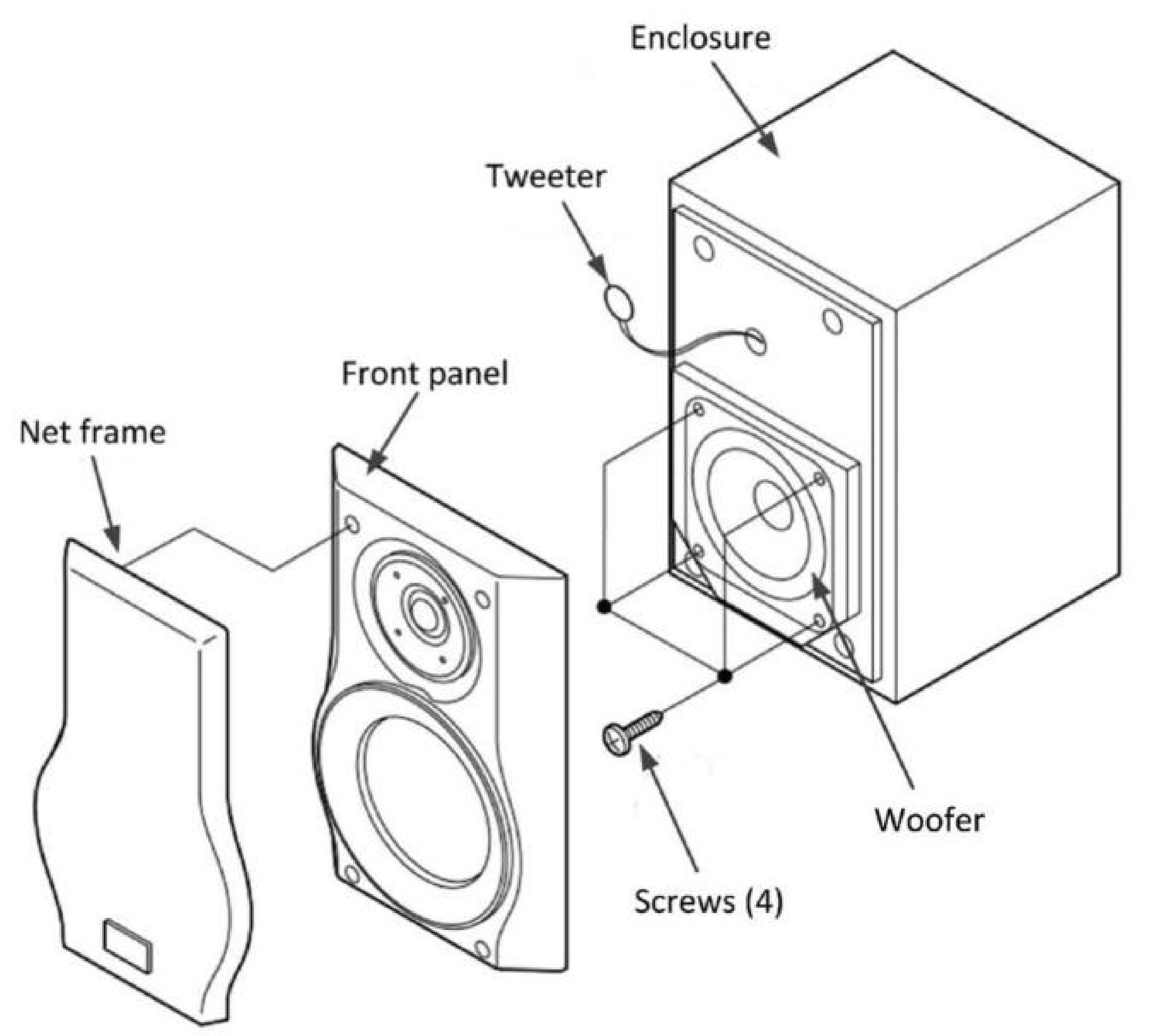
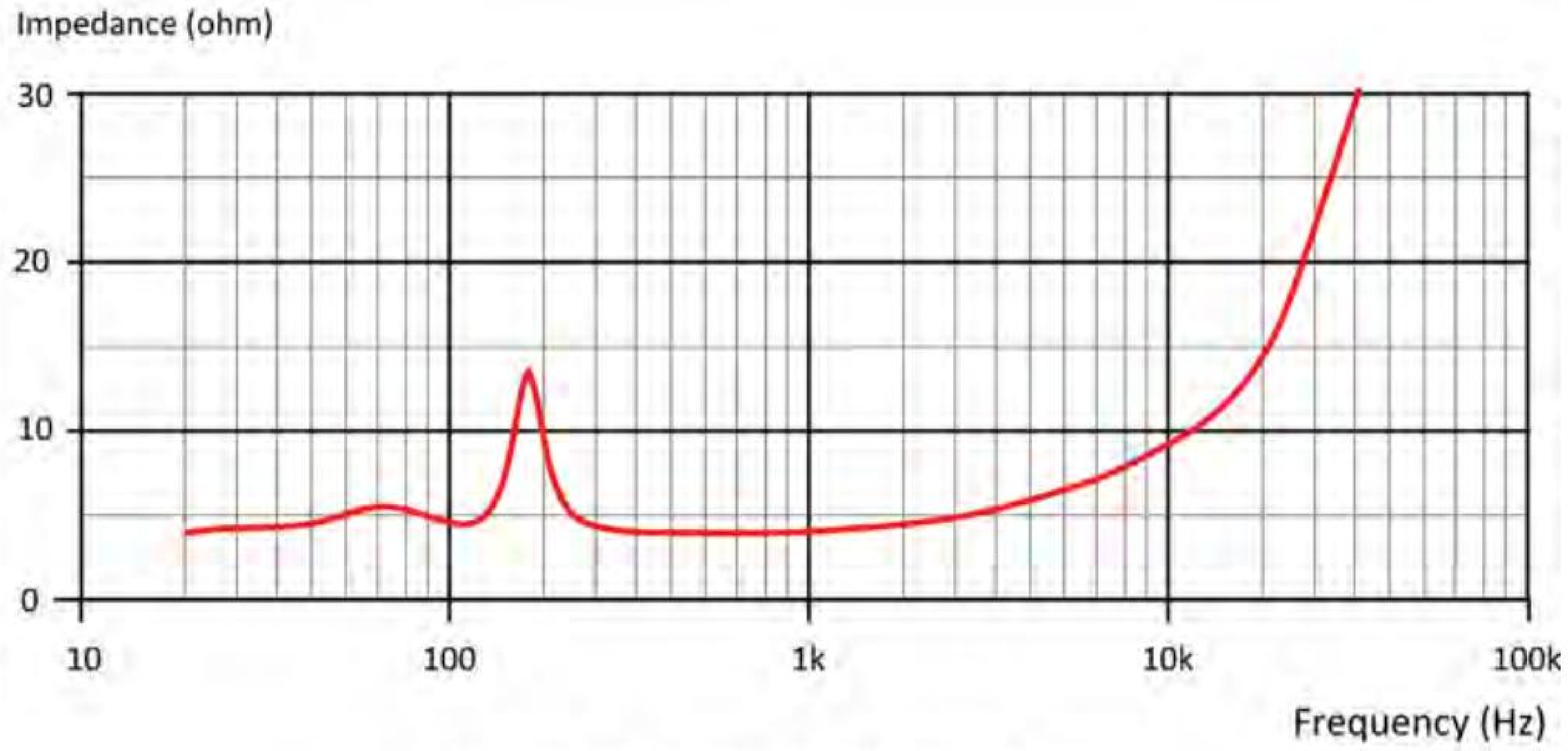
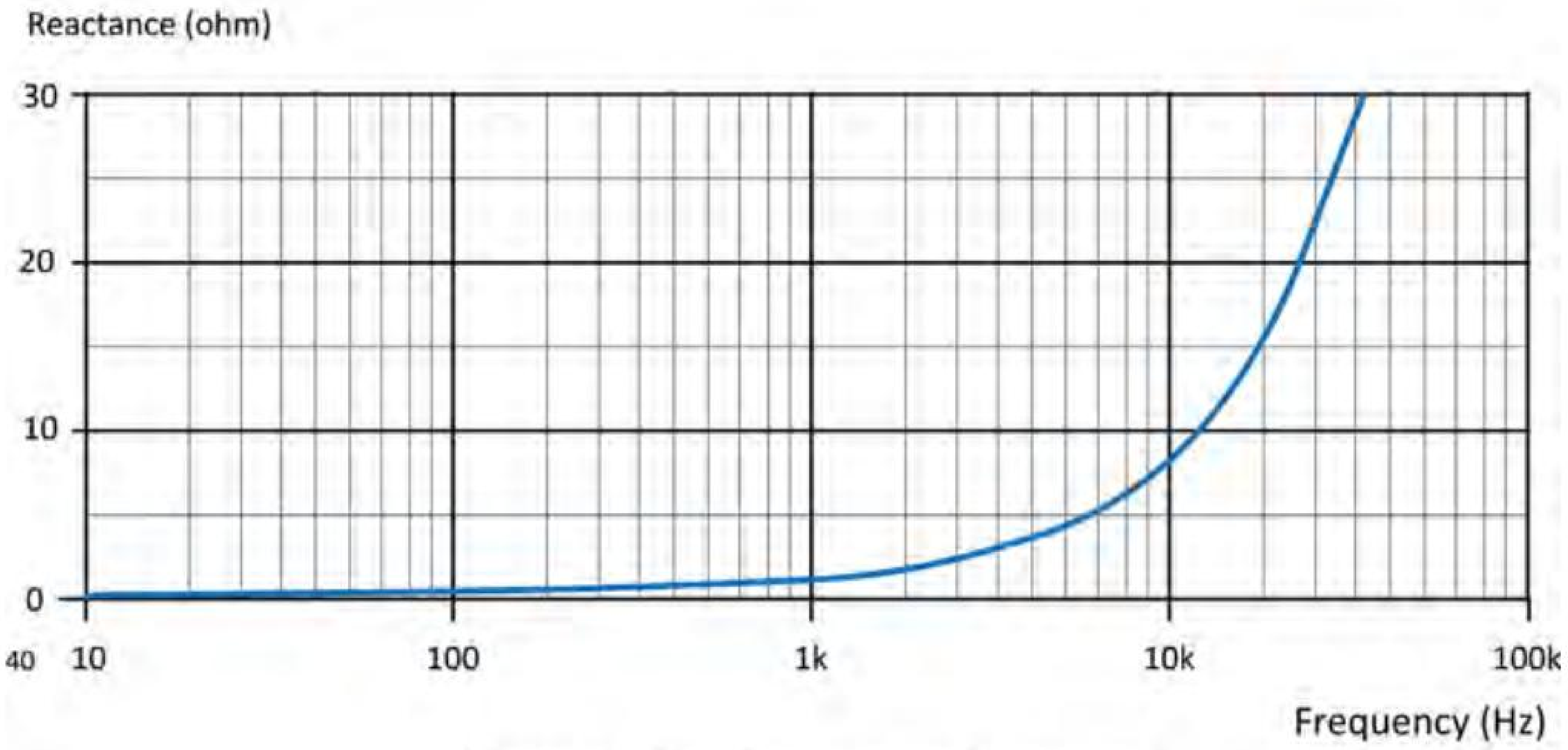


Fig.9.1. Construction of Sharp XL-60H loudspeaker system



(a) Loudspeaker impedance versus frequency



(b) Voice coil reactance versus frequency

Fig.9.2. Impedance/frequency characteristic for the Sharp XL-60H loudspeaker system

nature of the voice coil which, as mentioned earlier, was measured at 130µH. Fig.9.2b shows how this reactance varies with frequency. You should note that this curve rises from zero to about 8Ω at 10kHz and closely resembles the curve shown in Fig.9.2a. In practical terms, the upshot of this is that, due to the increase in impedance the current (and therefore also the power delivered to the loudspeaker) will become progressively reduced at frequencies above 4kHz. This problem can be partly offset by the addition of a tweeter – a speaker dedicated to higher frequencies – which helps to increase the sound pressure at high frequencies.

Equivalent circuit of a loudspeaker

Fig.9.3 shows the evolution of a loudspeaker as a load. In Fig.9.3b we have simply replaced the loudspeaker by a resistor having a DC resistance that is the same as the nominal impedance of the loudspeaker. This, as we mentioned earlier, is far from satisfactory in terms of modelling a loudspeaker's behaviour.

Fig.9.3c shows an improved equivalent circuit in which we've added the voice coil inductance. The impedance can be calculated at any frequency for this circuit using:

$$Z = \sqrt{R^2 + X_L^2} = \sqrt{R^2 + (2\pi f L)^2}$$

Fig.9.3d takes our equivalent circuit one step further. The circuit comprises a series network of R1 (the voice coil resistance), L1 (the voice coil inductance), and a parallel network comprising L2, C1, and R2. These three components are not 'real' physical components, but we have shown them here to account for the resonant peak in the impedance characteristic that we discussed earlier. Fig.9.4 shows the complete equivalent circuit of our Sharp XL-60H 40W loudspeaker with the values that we obtained by measurement and calculation.

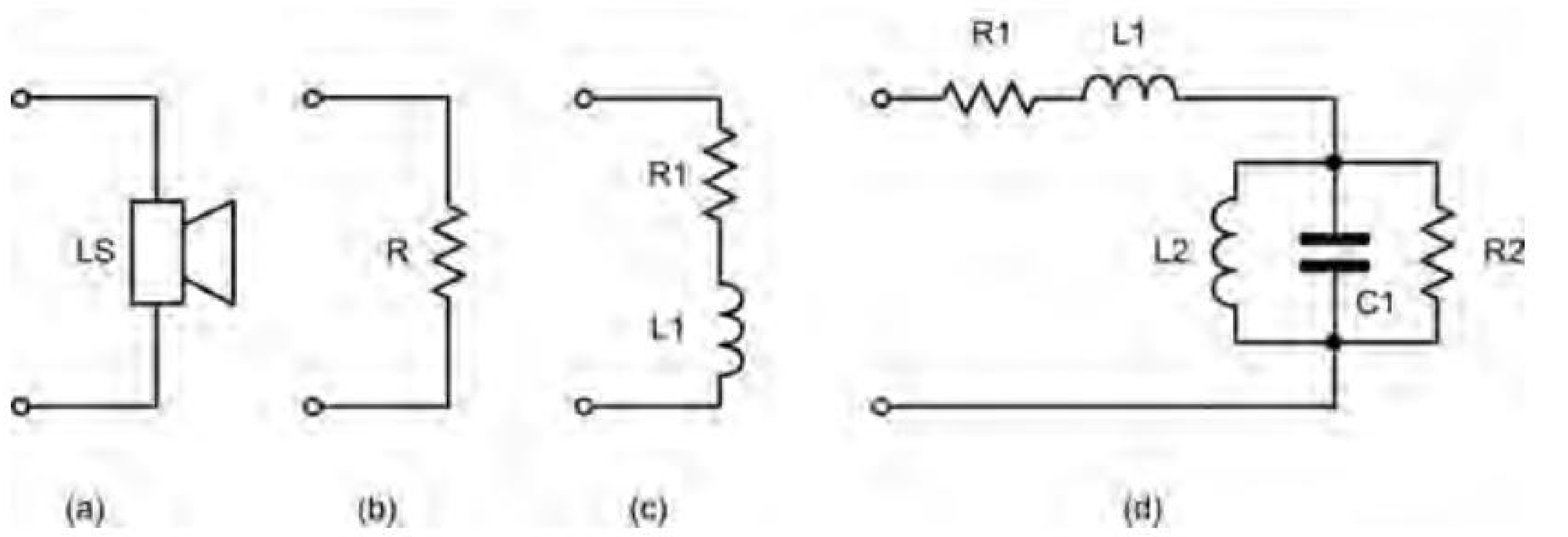


Fig.9.3. Evolution of a loudspeaker equivalent circuit

Zobel networks

Fortunately, the steady increase in impedance of a loudspeaker that results from the inductance of its voice coil can be counteracted by nothing more than a simple C-R network connected in parallel with the driver. This network is designed so that it exhibits an impedance which is equal and opposite to that of the voice coil. The result is a load that appears to be almost purely resistive at all frequencies. This clever idea came from the work of Otto Zobel, an engineer who was employed by the American telecommunications company, AT&T. A French electrical engineer, Paul Boucherot, who was then working for the French railway company, Chemin de Fer du Nord, had a similar idea, although his application was intended more for electrical power distribution than for use in the audio/communications sector.

Fig.9.5 shows a Zobel network connected in parallel with a loudspeaker. The values required for *RZ* and *CZ* (which essentially balance the 'bridge' circuit) can be calculated from:

$$RZ = R1$$

and

$$CZ = \frac{L1}{R1^2}$$

Multiple speaker systems

Good quality loudspeaker systems usually comprise several speakers within each speaker enclosure, and these are dedicated to different frequency ranges. Low-frequency driver units (woofers) generally cover from about 20Hz to 200Hz, while mid-range and high-frequency drivers (tweeters) usually cover the frequency range 200Hz to 20kHz. It then becomes necessary to separate the audio range into different bands so that each of the speakers is presented with an appropriate range of signals. This is usually achieved with the use of a 'cross-over network' of passive components arranged as simple second- or third-order L-C filters. Fig.9.6 shows how this works in the case of a dual loudspeaker system.

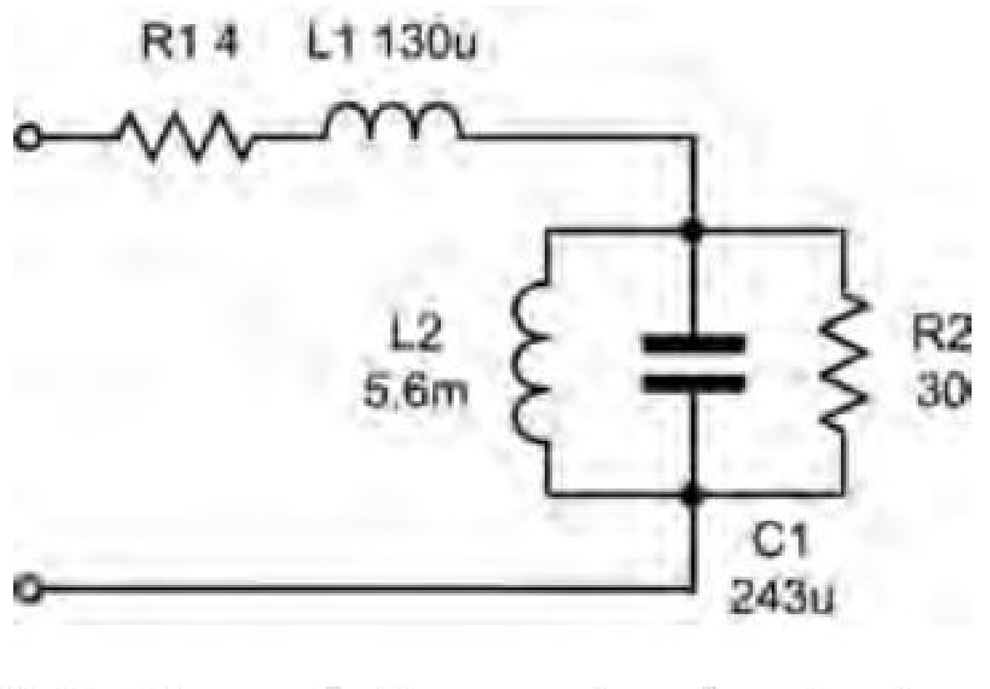


Fig.9.4. Complete equivalent circuit of our test loudspeaker system

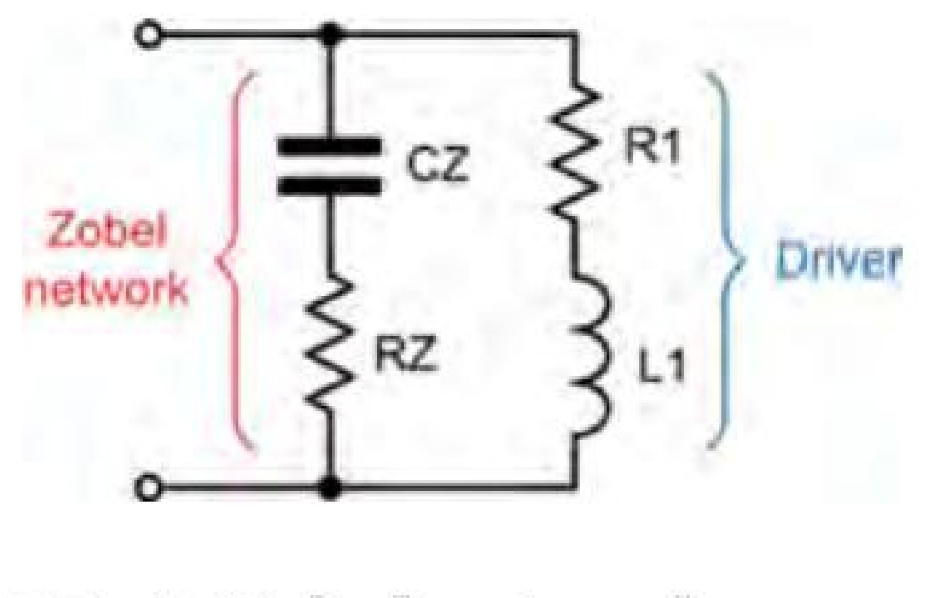


Fig.9.5. A Zobel network connected to a driver

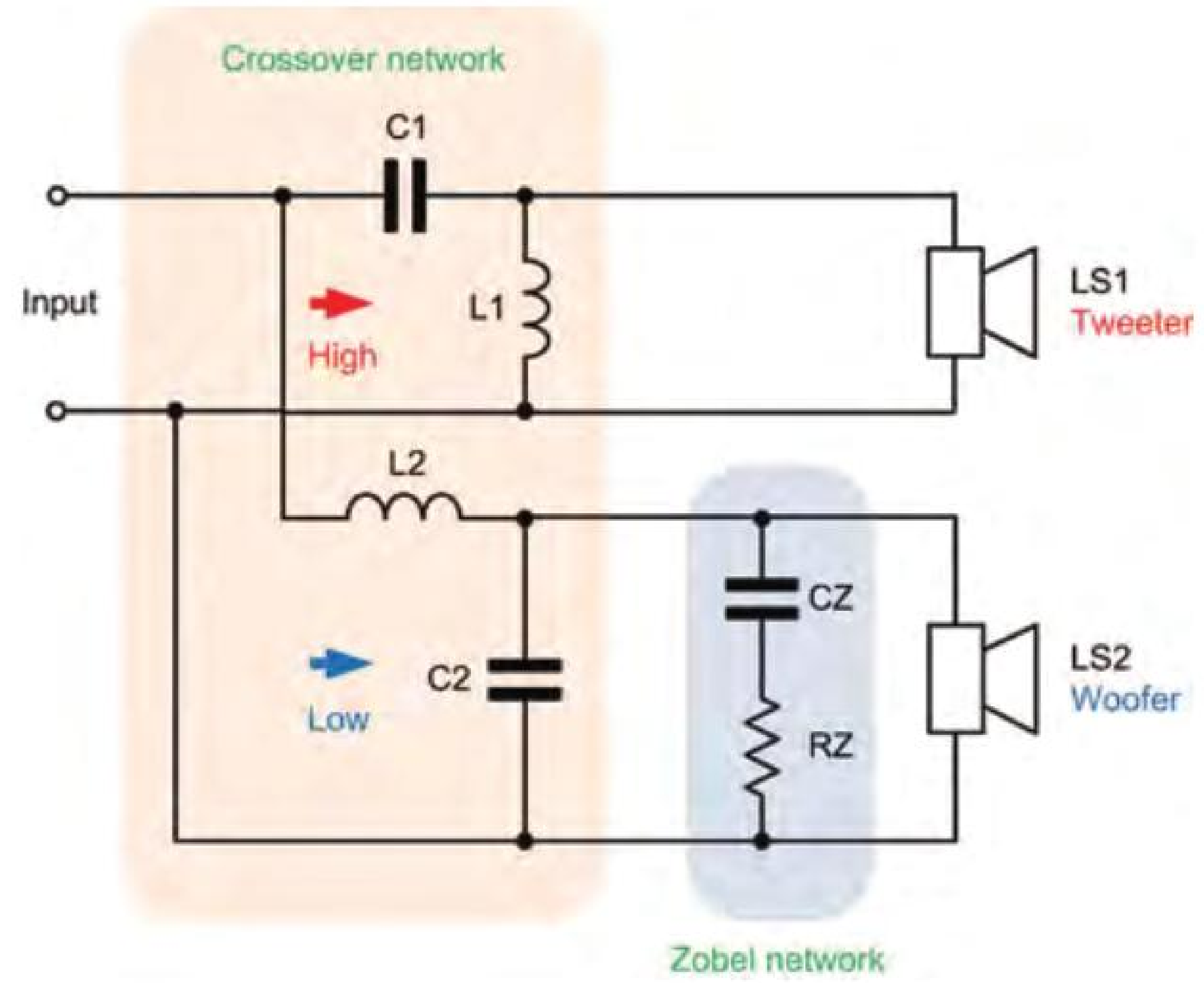

Crossover network
C1
Input
High
L1
LS1
Tweeter
L2
CZ
Low
C2
RZ
LS2
Woofer
Zobel network

Fig.9.6. A typical loudspeaker system with high- and low-range units and with Zobel and cross-over networks fitted

A second-order high-pass filter, C1/L1, is interspersed between the amplifier output and the high-range unit, LS1, while a second-order low-pass filter, L2/C2, is interspersed between the output and the low-range unit, LS2.

The design of the filters used in a cross-over network assumes that the loudspeaker (particularly the low-frequency driver) is well-behaved and has a constant impedance over the frequency range. As we've just shown, this is never the case, so one or more Zobel networks are invariably added, as shown in Fig.9.6. Note that *CZ* and *RZ* are fitted directly across the loudspeaker, not as part of the amplifier circuitry.

Using Tina Design Suite to model a loudspeaker

The performance of a loudspeaker, with and without a Zobel network fitted, can be easily modelled using Tina Design Suite, as shown in Fig.9.7. Note that Tina cannot plot impedance directly, so instead you use Tina's Signal Analyzer tool to plot voltage against frequency. In Fig.9.8 we plotted the voltage/frequency characteristic when the loudspeaker (without an added Zobel network) was supplied with 1V from a source having an internal impedance of 4Ω. In this graph, a voltage of 500mV corresponds to a load impedance of 4Ω. This is the case at 10Hz and 1kHz but, as expected, the impedance rises to a peak at around 150Hz and also rises progressively towards the high end of the spectrum.

In Fig.9.9 we used Tina's Signal Analyzer to plot the response of the system with a Zobel network present. In this case, *RZ* was 4Ω and *CZ* was 8µF, but please feel free to check these values by calculation (use the formula that we showed earlier). Note that the resonant peak is still present but, above a few hundred Hz the impedance does remain substantially constant at almost exactly 4Ω, clearly demonstrating the usefulness of Zobel's work.

Improving stability

Earlier in this series we showed how the stability of an amplifier can be improved by adding negative feedback. But, as we've just demonstrated, there's always a degree of uncertainty concerning the load to which an amplifier is connected. This load can often have significant reactance and because of this, additional precautions are usually necessary to ensure that an amplifier remains unconditionally stable.

A typical output network for a high-power audio amplifier is shown in Fig.9.10. R1 and R2 are low-value emitter resistors that add some series current feedback to help maintain the operating conditions of TR1 and TR2. They also help ensure that the amplifier has a measure of protection against a short circuit at its output. R1 and R2 must be suitably rated. In the case of an amplifier rated at a fairly modest 10W into 4Ω, an RMS output current of around 1.5A will be produced and so R1 and R2 will need to be rated for a power of at least 1W.

The series network formed by R3 and C1 connected in parallel with the output is not, as it might first appear, a Zobel network for correcting the rising impedance characteristic of the loudspeaker (see earlier). Instead, it ensures that the amplifier is always loaded at frequencies outside the normal audio frequency range. This is

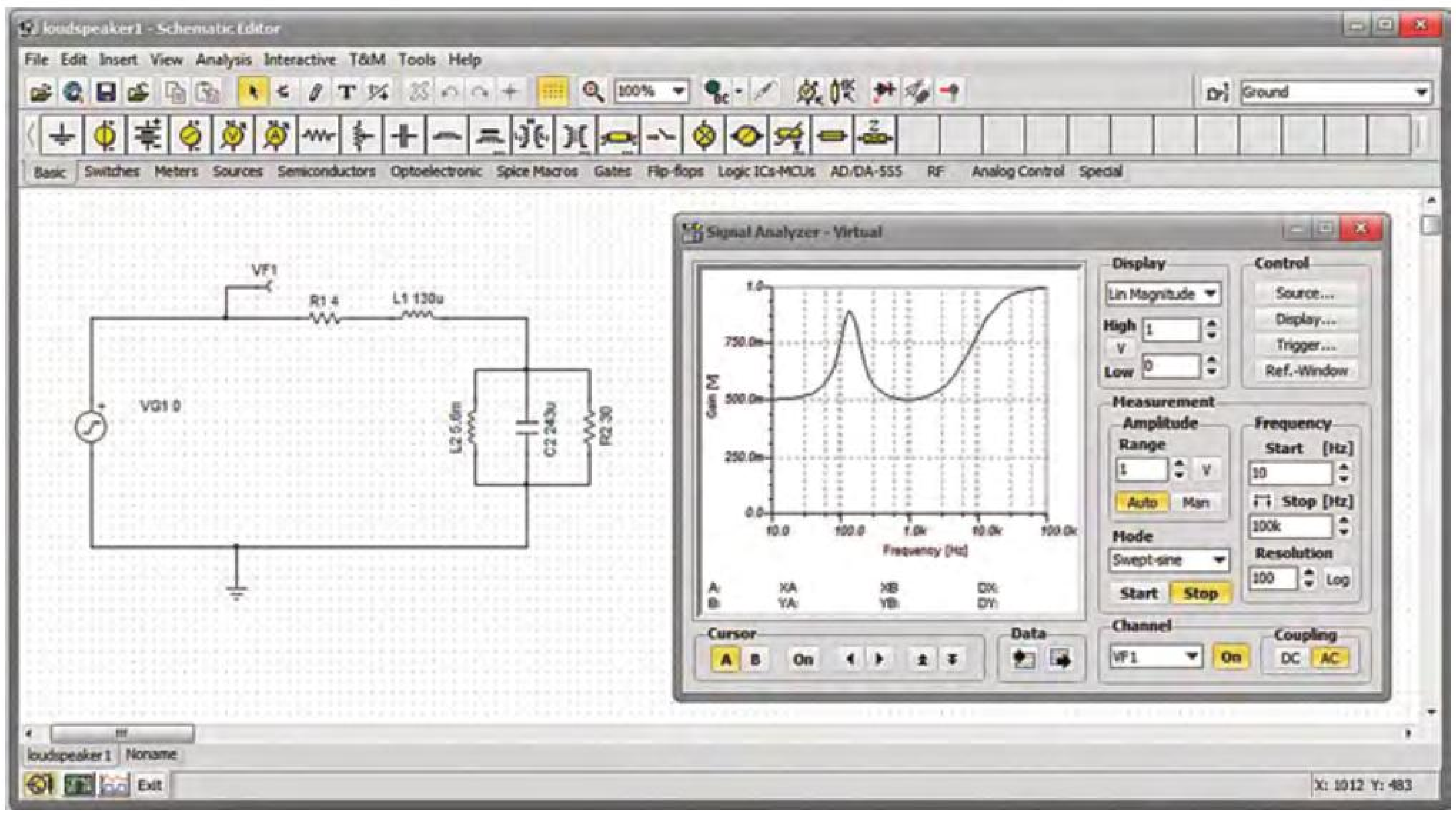

loudspeaker1 - Schematic Editor
File Edit Insert View Analysis Interactive T&M Tools Help
100%
Ground
Basic Switches Meters Sources Semiconductors Optoelectronic Spice Macros Gates Flip-flops Logic ICs-MCUs AD/DA-555 RF Analog Control Special
VF1
R1 4
L1 130u
VG1 0
L2 5.6m
C2 243u
R2 30
Signal Analyzer - Virtual
1.0
750.0m
500.0m
250.0m
0.0
10.0
100.0
1.0k
10.0k
100.0k
Frequency [Hz]
Gain [V]
A: XA: XB: DX:
B: YA: YB: DY:
Display
Lin Magnitude
High 1 V
Low 0
Control
Source...
Display...
Trigger...
Ref.-Window
Measurement
Amplitude
Range
1 V
Auto Man
Frequency
Start [Hz]
10
Stop [Hz]
100k
Resolution
100 Log
Mode
Swept-sine
Start Stop
Cursor
A B On
Data
Channel
VF1 On
Coupling
DC AC
loudspeaker1 Noname
Exit
X: 1012 Y: 483

Fig.9.7. Using Tina Design Suite to model the behavior of the Sharp XL-60H loudspeaker system

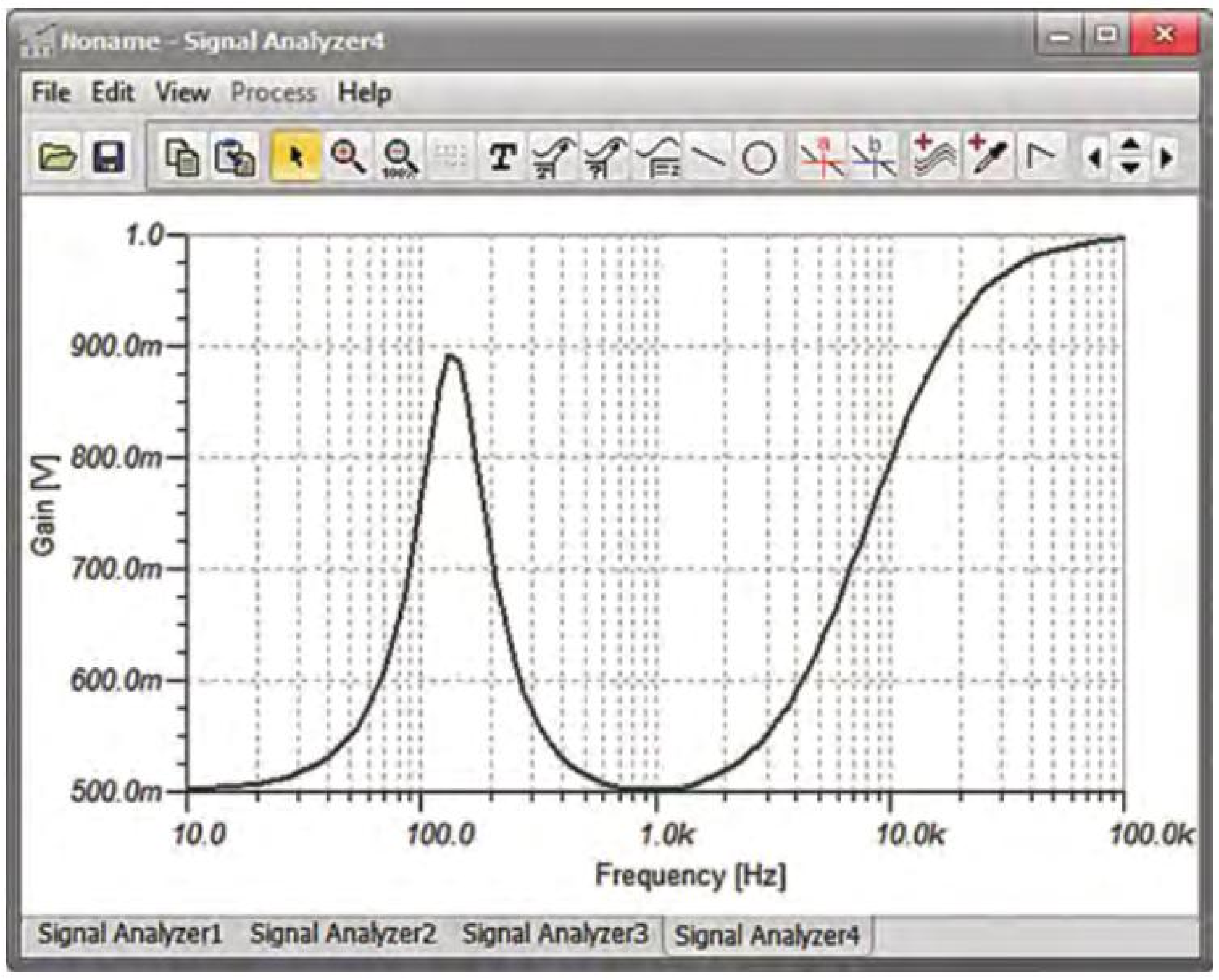


Fig.9.8. Using Tina's Signal Analyzer to plot the network response without a Zobel network fitted

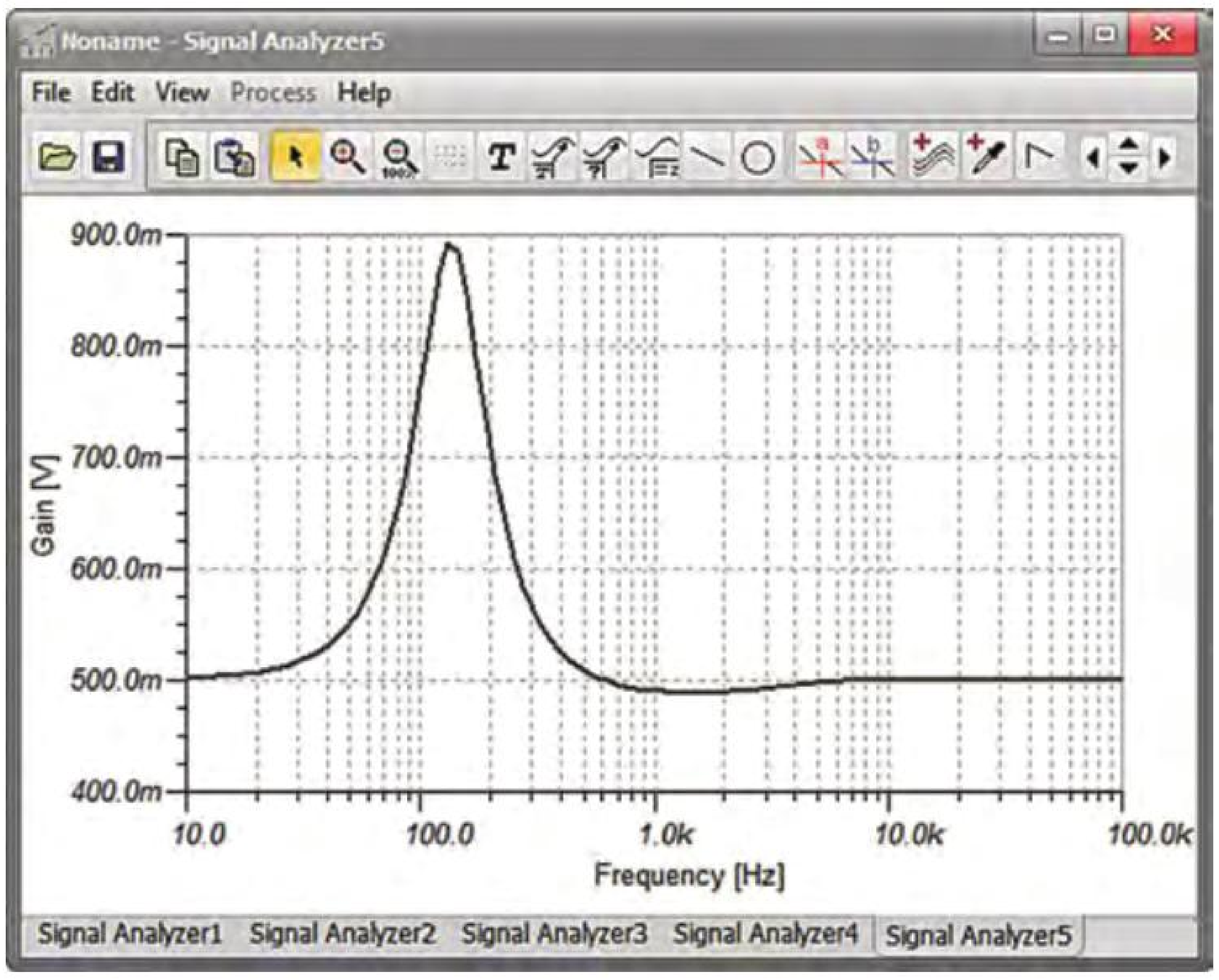


Fig.9.9. Using Tina's Signal Analyzer to plot the network with a Zobel network fitted

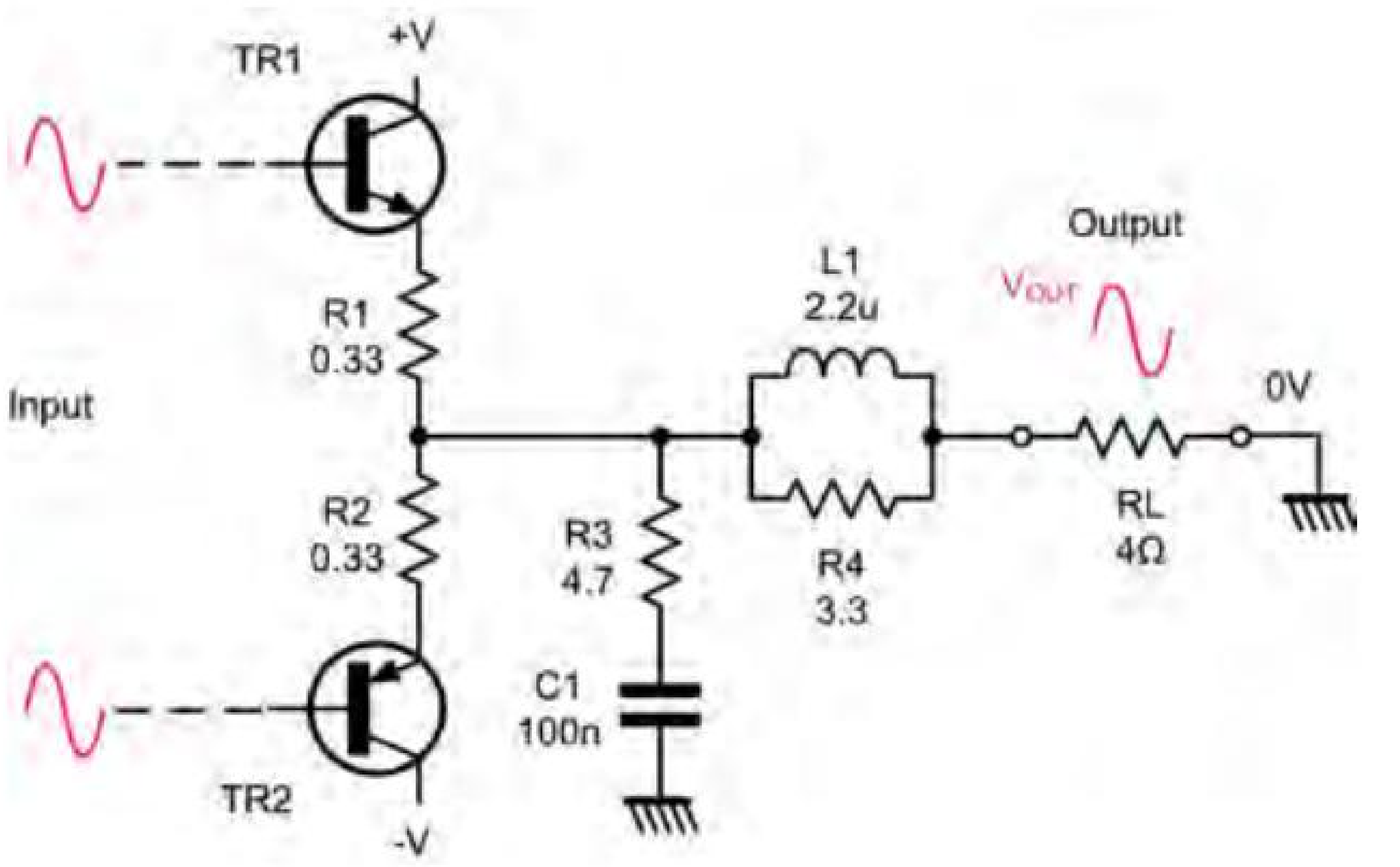


Fig.9.10. Output network showing typical component values

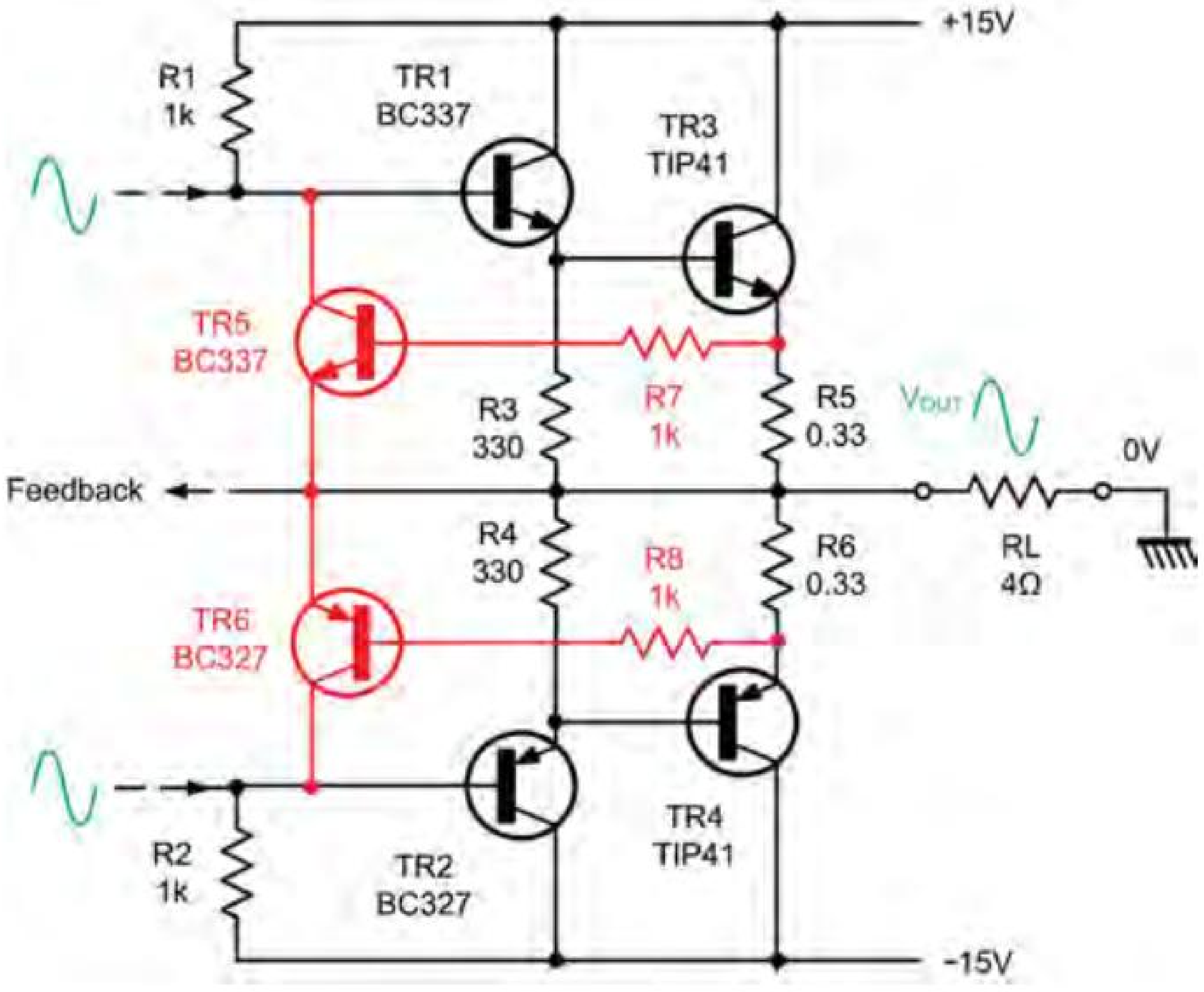


Fig.9.11. Typical complementary emitter-follower output stage incorporating over-current protection

sometimes necessary due to an inherent susceptibility of emitter-follower output stages to oscillate at frequencies well outside the audio spectrum. With the values shown in Fig.9.8 the network ensures that the amplifier sees a load of around 5Ω at a frequency of 1MHz. In addition, the parallel network formed by L1 and R4 connected in series with the output helps to increase the impedance of the output at high frequencies (the reactance of L1 is about 14Ω at 1MHz).

Over-current and short-circuit protection

With high power amplifiers and due to the vagaries of load impedances and, more importantly, the possibility of an inadvertent short circuit at the output, it is desirable to incorporate some form of over-current protection in the output stage. A typical arrangement is shown in Fig.9.11. TR1, TR2, TR3 and TR4 form a complementary cascaded emitter-follower output stage (see last month). The voltage across the two emitter resistors, R5 and R6, will be directly proportional to the emitter currents of TR3 and TR4 respectively. If the voltage dropped across R5 or R6 should ever exceed the base-emitter voltage threshold (around 0.6V) for TR5 and TR6, the respective transistor will conduct heavily, effectively clamping the base bias voltage applied to TR1 or TR2. This, in turn, will have the effect of limiting the output current to about 2A with the values shown in Fig.9.11. Short-circuiting the output terminals under driven conditions will then only have the effect of limiting the output current and voltage swing (at which point the output waveform will become severely clipped).

Thermal protection

Earlier in Part 7 (August 2015, *EPE*) we discussed heat and heat dissipation, and described some practical heatsink arrangements that can be used with audio power amplifiers, the aim being to prolong component life and avoid early failure. A properly designed heatsink mounted in a ventilated enclosure will help limit the temperature of the output devices but, due to the vagaries of the environment in which an amplifier operates (particularly as regards ambient temperature), it may become necessary to incorporate an electronic means of limiting the rise in temperature experienced by the output transistors. We can do this easily by incorporating *thermal feedback* into our design. The best place to apply such feedback is in the stage that sets the bias for the output devices.

Thermal feedback is based on sensing the temperature of the output devices in order to regulate the amount of bias voltage applied to them. There are several ways of doing this, but one of the most effective makes use of the positive temperature coefficient that is exhibited by conventional bipolar junction transistors. When the junction temperature increases the base-emitter threshold voltage, V_{BE}, will fall at a rate of about 2mV per °C.

As an example, a BC847 transistor operating with a collector current of 10mA will exhibit a base-emitter threshold voltage of 700mV at 25°C but, at 75°C this will have fallen to around 600mV (see Fig.9.12).

As temperature increases, the falling threshold voltage will produce an *increase* in base current. If this sounds a little counterintuitive it is worth recalling that the base connection of a V_{BE} multiplier is fed from a potential divider (see Fig.9.13). If the base-emitter threshold voltage falls, more base current will flow and, in turn, this produces a corresponding increase in collector current. As a result, the collector-emitter voltage (V_{BIAS} in Fig.9.13) will fall. The reduction in V_{BIAS} will result in a decrease in base bias current applied to both the driver and output transistors and, as a result, the power dissipation will be reduced and the temperature of the output devices will fall.

The device used for the V_{BE} multiplier stage should be mounted in very close proximity to one or both of the output devices (ie, it should have close thermal coupling). In some cases the sensing device is attached directly to the case of one of the output transistors, but in others it is simply clamped onto the heatsink between the final output pair. Last, it

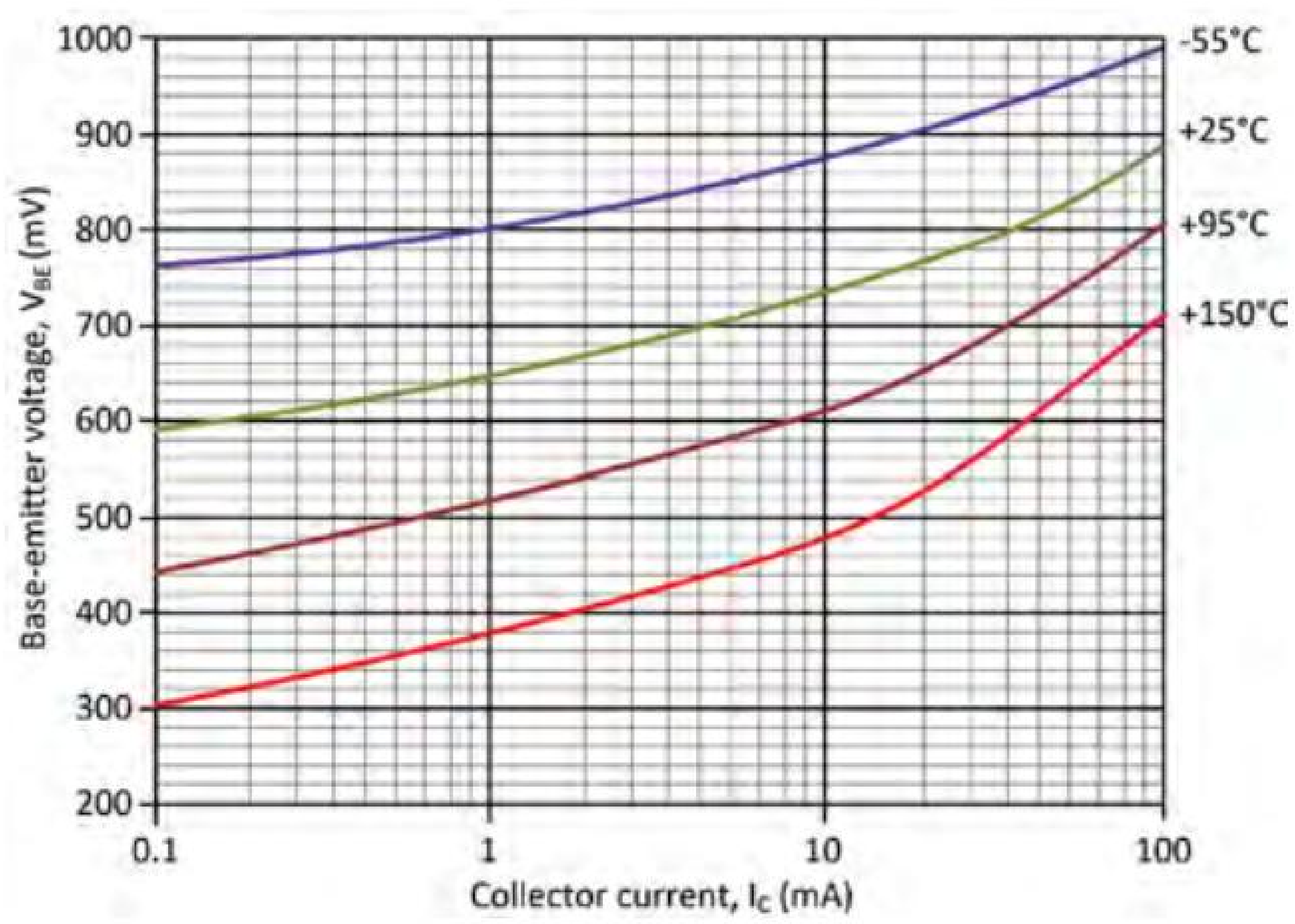


Fig.9.12. Base-emitter voltage as a function of temperature for a collector current of 10mA

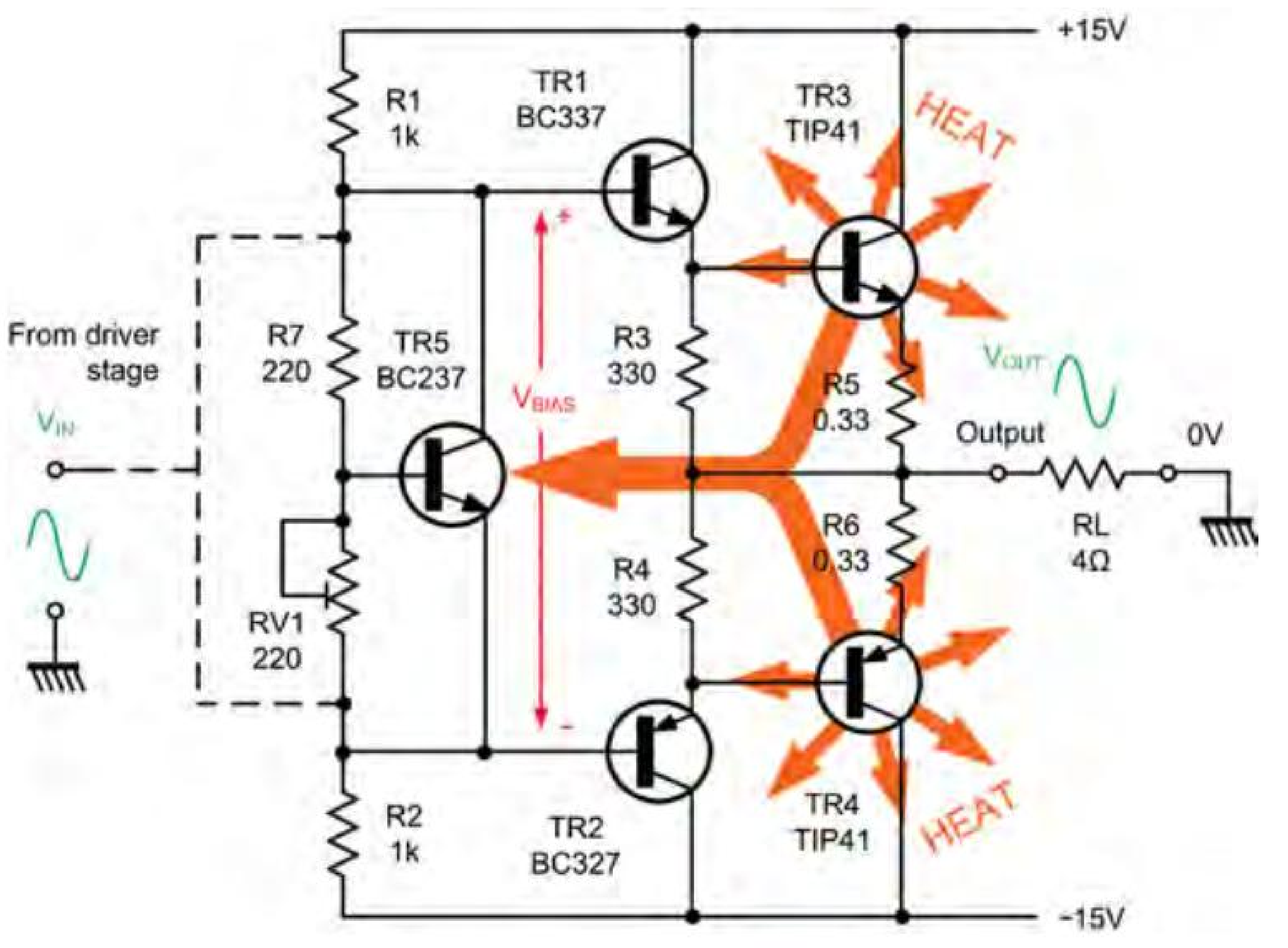


Fig.9.13. V_{BE} multiplier with thermal feedback from output stage

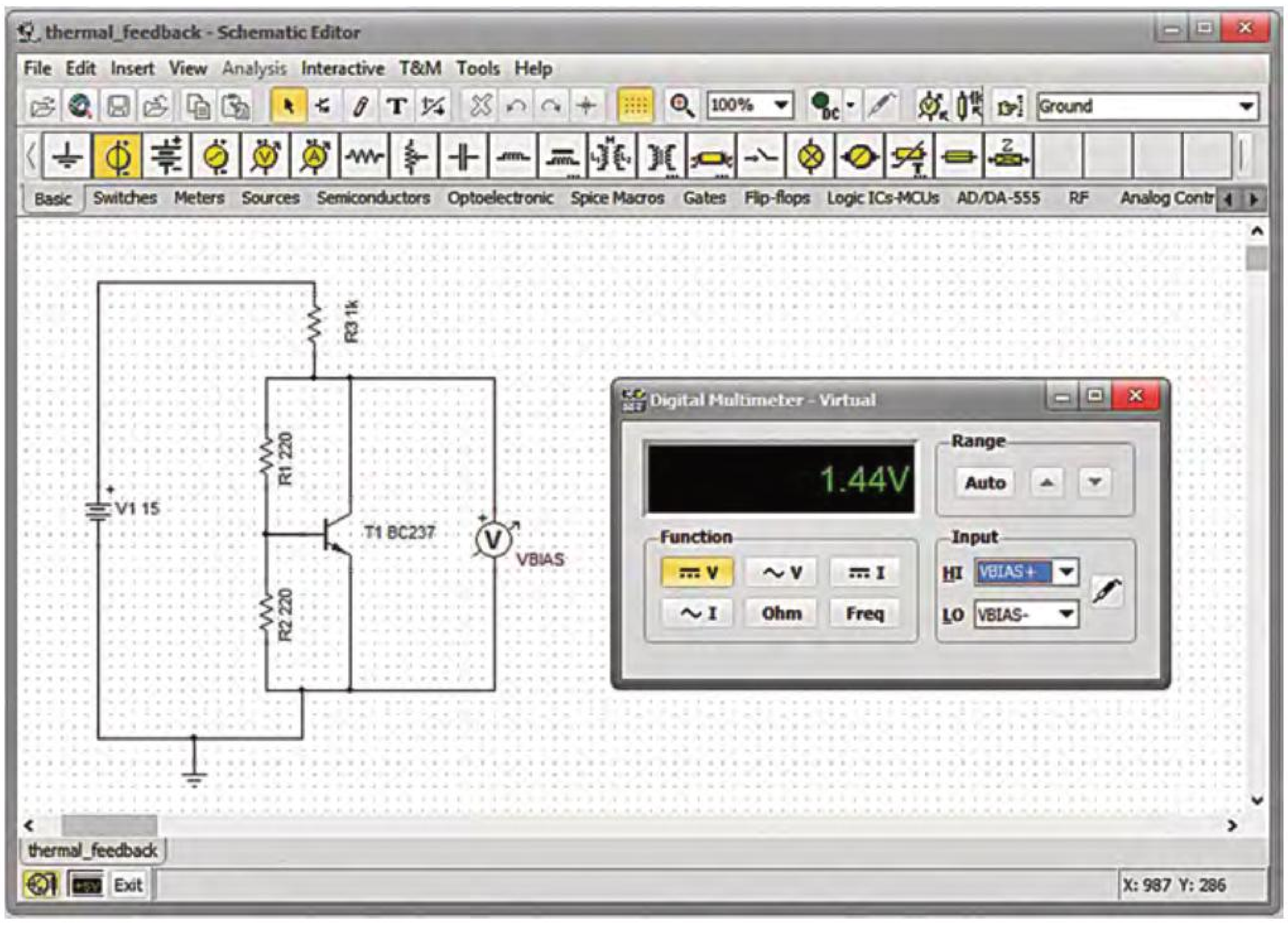
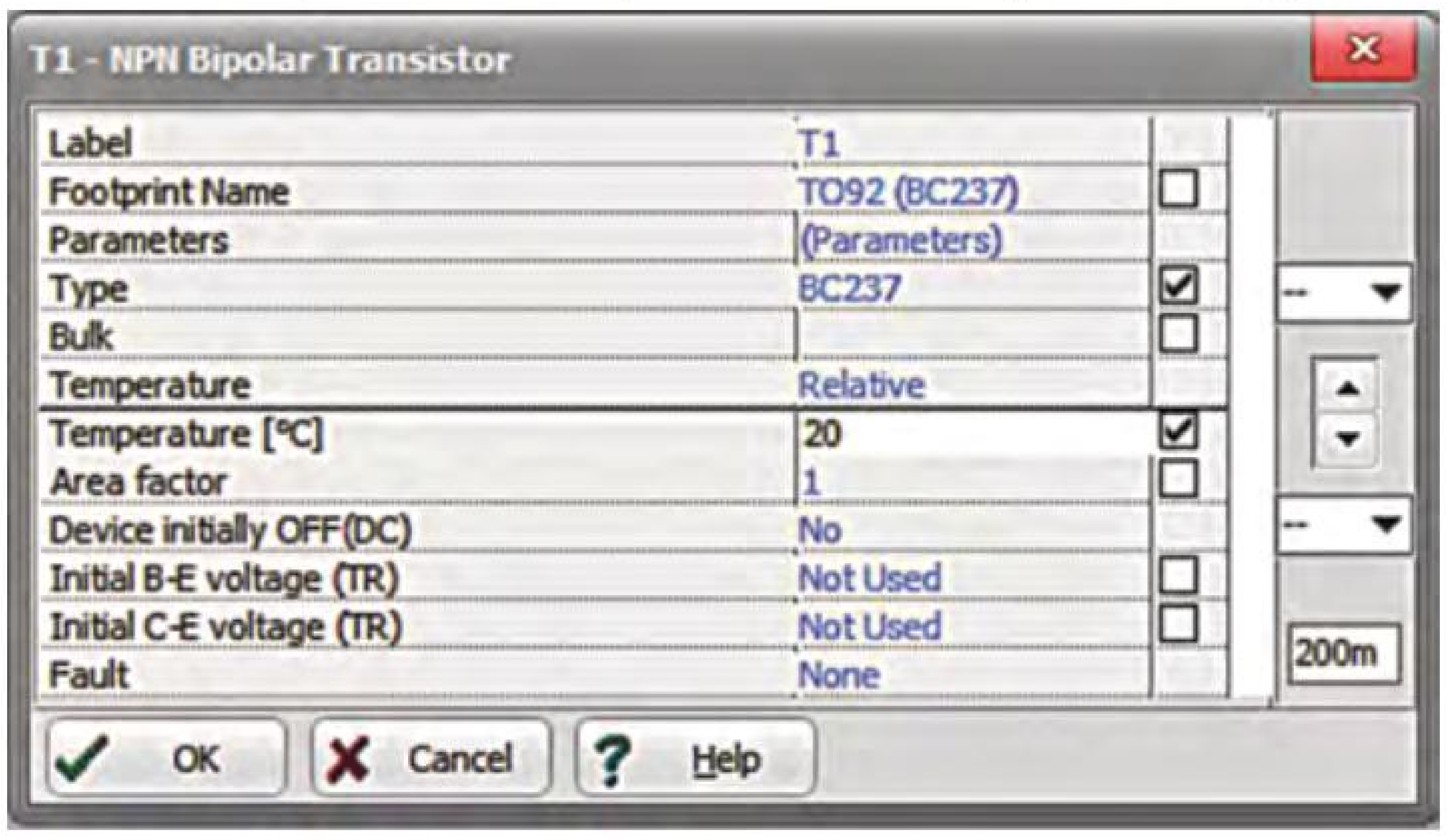

thermal_feedback - Schematic Editor
File Edit Insert View Analysis Interactive T&M Tools Help
100%
Ground
Basic Switches Meters Sources Semiconductors Optoelectronic Spice Macros Gates Flip-flops Logic ICs-MCUs AD/DA-555 RF Analog Contr
R3 1k
R1 220
R2 220
V1 15
T1 BC237
VBIAS
Digital Multimeter - Virtual
1.44V
Range
Auto
Function
V
V
I
I
Ohm
Freq
Input
HI VBIAS+
LO VBIAS-
thermal_feedback
Exit
X: 987 Y: 286

Fig.9.14. Testing the VBE multiplier at different temperatures



Label	T1	
Footprint Name	TO92 (BC237)	☐
Parameters	(Parameters)	
Type	BC237	☑
Bulk		☐
Temperature	Relative	
Temperature [°C]	20	☑
Area factor	1	☐
Device initially OFF(DC)	No	
Initial B-E voltage (TR)	Not Used	☐
Initial C-E voltage (TR)	Not Used	☐
Fault	None	

Fig.9.15. Setting transistor parameters and ambient temperature in Tina Design Suite

is worth noting that in Fig.9.13, pre-set potentiometer RV1 has been included in the lower section of the V_{BE} multiplier's potential divider. This helps to ensure that, in the event of this component failing in an open-circuit condition, the output stage is biased completely off.

You can easily check the operation of a V_{BE} multiplier as a temperature sensing device using Tina Design Suite, as shown in Fig.9.14. We used Tina's virtual multimeter to display the output voltage, V_{BIAS}. Fig.9.15 shows Tina's transistor parameter dialogue in which the ambient temperature can be set. Note the tick in the checkbox which allows the scroll arrows (on the right) to be used for adjusting the ambient temperature. Finally, Fig.9.16 shows the results of our investigation plotted as a graph. This clearly shows how the bias voltage falls from about 1.44V at 20°C to around 1.25V at 80°C.

Get Real: A high-performance 10W audio power amplifier

Our final *Get Real* project is designed to satisfy the need for a simple, low-cost power amplifier that can be used with other modules in this series to realise a complete modular audio system. When we originally discussed this series with editor Matt Pulzer, I suggested that it would be possible to produce a simple low-cost audio amplifier based on discrete components that would out-perform existing designs based on integrated circuits. This is the culmination of our *Get Real* practical projects; an amplifier capable of excellent performance at a reasonable cost. Our original minimum design specification for the amplifier is given in Table 9.1.

Circuit description

The complete circuit of the high-quality audio amplifier is shown in Fig.9.17. The amplifier uses eight commonly available low-cost silicon transistors. The input stage, comprising TR1 and TR2, is connected in differential configuration with a constant current source (the 'tail') formed by TR3 and associated components. This device supplies the combined emitter current for the differential pair, TR1 and TR2. A PNP device, TR4, acts as a common-emitter driver stage, while TR5 and associated components connected in the collector circuit of TR4 act as a V_{BE} multiplier, which defines and stabilises the bias applied to the output stage formed by TR6, TR7 and TR8. The output stage comprises a quasi-complementary configuration in which a Sziklai compound feedback pair, TR7 and TR8, replaces the single PNP device that might otherwise have been fitted as a complement to the NPN emitter-follower stage, TR6. To help you understand the circuit, Fig.9.17 shows the five functional stages shaded together with the negative feedback path from the output stage back to the differential amplifier.

Gain and frequency response

The input signal is coupled into the pre-amplifier by means of C1 but, apart from this, the amplifier is entirely DC coupled. This ensures excellent low-frequency response. DC negative feedback from the output is provided by means of R6. This helps to regulate the DC conditions within the amplifier. AC negative feedback is used to define the overall voltage gain and frequency response. The overall mid-band voltage gain is approximately given by:

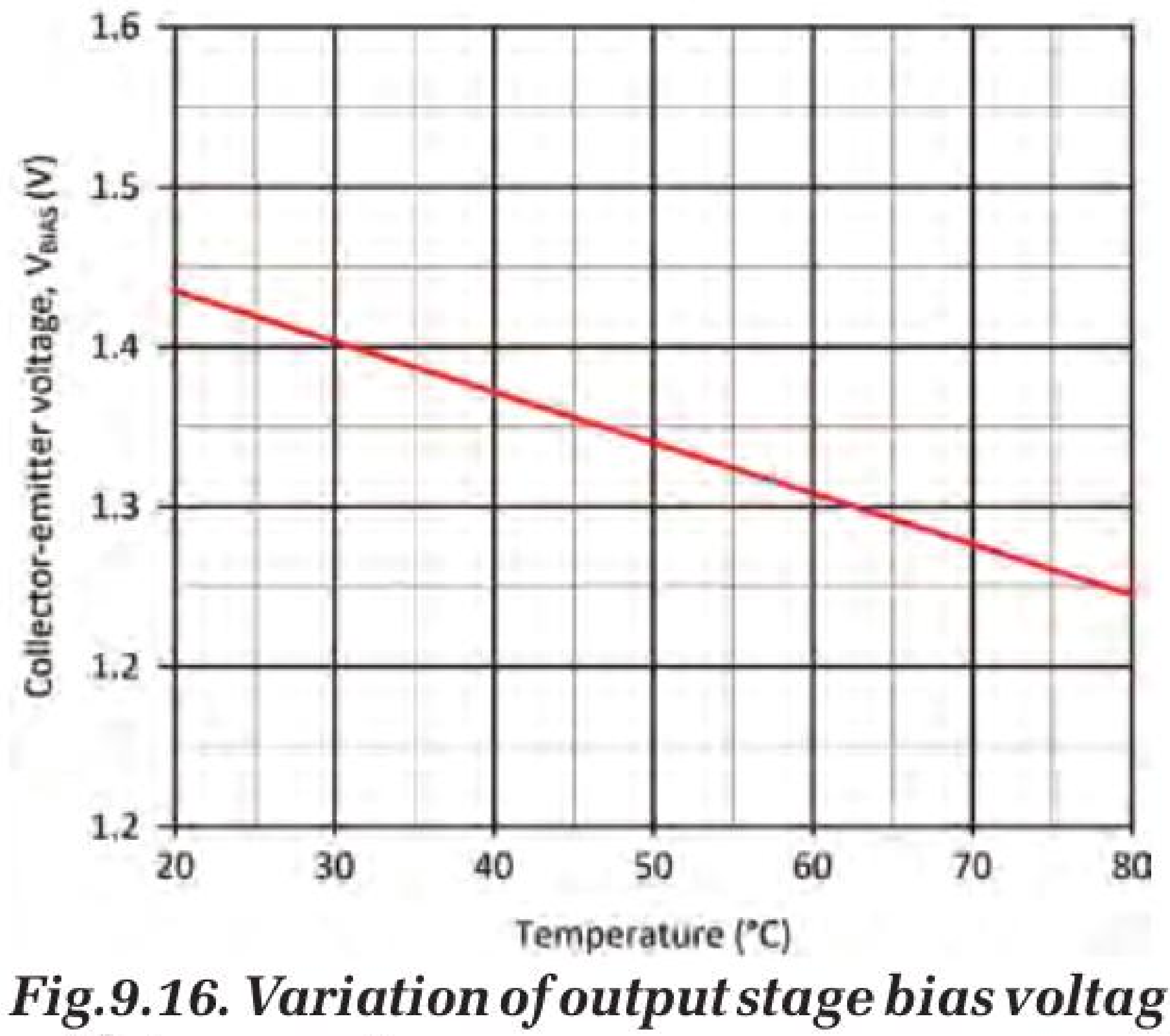


Fig.9.16. Variation of output stage bias voltage with temperature

Table 9.1 Outline design specification for the high-quality 10W audio amplifier

Device	Emitter
Voltage gain	10 (approx.)
Frequency response	Better than 10Hz to 20kHz at –1dB
Input impedance	50kΩ (approx.)
Output load impedance	3Ω to 15Ω (4Ω nominal)
Output power	10W into 4Ω at 1kHz at less than 1% THD
Phase shift	0° at 1kHz (ie, output non-inverted)
Supply voltage	12V to 18V (15V nominal)
Supply current	50mA (no signal), 1A (max.)
Distortion	Better than 0.5% THD at 5W output into 4Ω

$$A_V = 1 + \frac{R6}{R5} = 1 + \frac{100\times10^3}{10\times10^3} = 11$$

The lower-cut-off frequency is largely defined by the parallel network formed by R6 and C3, while the upper cut-off frequency is mainly defined by the series network formed by R5 and C2. The overall mid-band voltage gain, A_V, is approximately given by:

$$A_V = 1 + \frac{R6}{R5} = 1 + \frac{100\times10^3}{10\times10^3} = 11$$

The lower cut-off frequency, f_1, is thus approximately:

$$f_1 = \frac{0.159}{C2\times R5} = \frac{0.159}{4.7\times10^{-6}\times10\times10^3} = 3.4$$

While the upper cut-off frequency, f_2, is thus approximately:

$$f_2 = \frac{0.159}{C3\times R6} = \frac{0.159}{47\times10^{-12}\times100\times10^3} = 34\text{ kHz}$$

Where necessary, an external volume control (for example, a good quality 100kΩ logarithmic-law carbon potentiometer) can be fitted at the input of the amplifier.

Bias adjustment

Two adjustments are required; RV1 sets the symmetry and allows the output voltage to be set to zero in the absence of a signal, while RV2 sets the standing bias current in the two output devices, TR6 and TR7. The correct adjustment of these two pre-set controls is crucial in determining the overall performance of the amplifier.

The setting of RV1 determines the collector current of the constant current source, TR3. This, in turn, impacts on the collector voltage of TR1. If this voltage falls, TR4 will conduct more heavily and, as a result, the voltage at the base of TR6 will rise. This, in turn, will cause the output voltage (at the junction of R10 and R11) to increase. Conversely, if TR4's collector voltage rises, TR4 will conduct less heavily and, as a result, the voltage at the base of TR6 will fall. This, in turn, will cause the output voltage (at the junction of R10 and R11) to fall.

The setting of RV2 determines the collector-emitter voltage of the V_{BE} multiplier formed by TR5 and associated components. This voltage also acts as the base bias voltage applied to TR6 and TR7. This voltage should be nominally around 1.2V to bias the two base-emitter junctions for linear operation but, as we explained earlier, the voltage will be dependent on the ambient temperature and thus will also provide some thermal feedback to protect the output stage. C5 simply ensures that the same signal voltage is applied (in-phase) to the base of TR6 and TR7.

Finally, it's worth noting that these two low-value resistors have been included to provide some protection for TR6 and TR8. As the current flowing in R10 and R11 increases, the base-emitter voltages for TR6 and TR7 will be correspondingly reduced. So, for example, if the emitter current in TR6 or TR8 was to increase, by say 50mA, the respective base bias voltage would be reduced by 50mV, opposing the original change and helping to stabilise the emitter current.

Supplies

The positive and negative supply rails are derived from an external supply of nominally ±15V. In practice, any symmetrical voltage supply of between ±12V and ±17V rated at 1.5A can be used, but the maximum output power will fall to about 8W with supplies of less than about ±13.5V. Decoupling of the positive and negative power rails is provided by C7 and C8 respectively. Note that these two components will effectively appear in parallel with the larger smoothing/reservoir capacitors present in the power supply (more of this next month) and so their use here is primarily to decouple signals that might otherwise appear on the supply rails.

Components

1 PCB, code 909 available from the *EPE PCB Service*, size 163mm × 56mm
2 PCB mounting 2-way terminal blocks
1 PCB mounting 3-way terminal block
2 Heatsinks (*see text*)

Resistors

1 100kΩ 0.25W (R1)
1 470Ω 0.25W (R2)
1 3.9kΩ 0.25W (R3)
3 100Ω 0.25W (R4, R7, R8)
1 10kΩ 0.25W (R5)
1 100kΩ 0.25W (R6)
1 560Ω 0.5W (R9)
2 1Ω 1W (R10 and R11)
1 4.7Ω 1W (R12)
1 500Ω miniature skeleton pre-set (RV1)
1 100Ω miniature skeleton pre-set (RV2)

Capacitors

1 1µF (C1)
1 4.7µF (C2)
1 47pF (C3)
1 220pF (C4)
1 220µF (C5)
1 100nF (C6)
2 10µF (C7, C8)

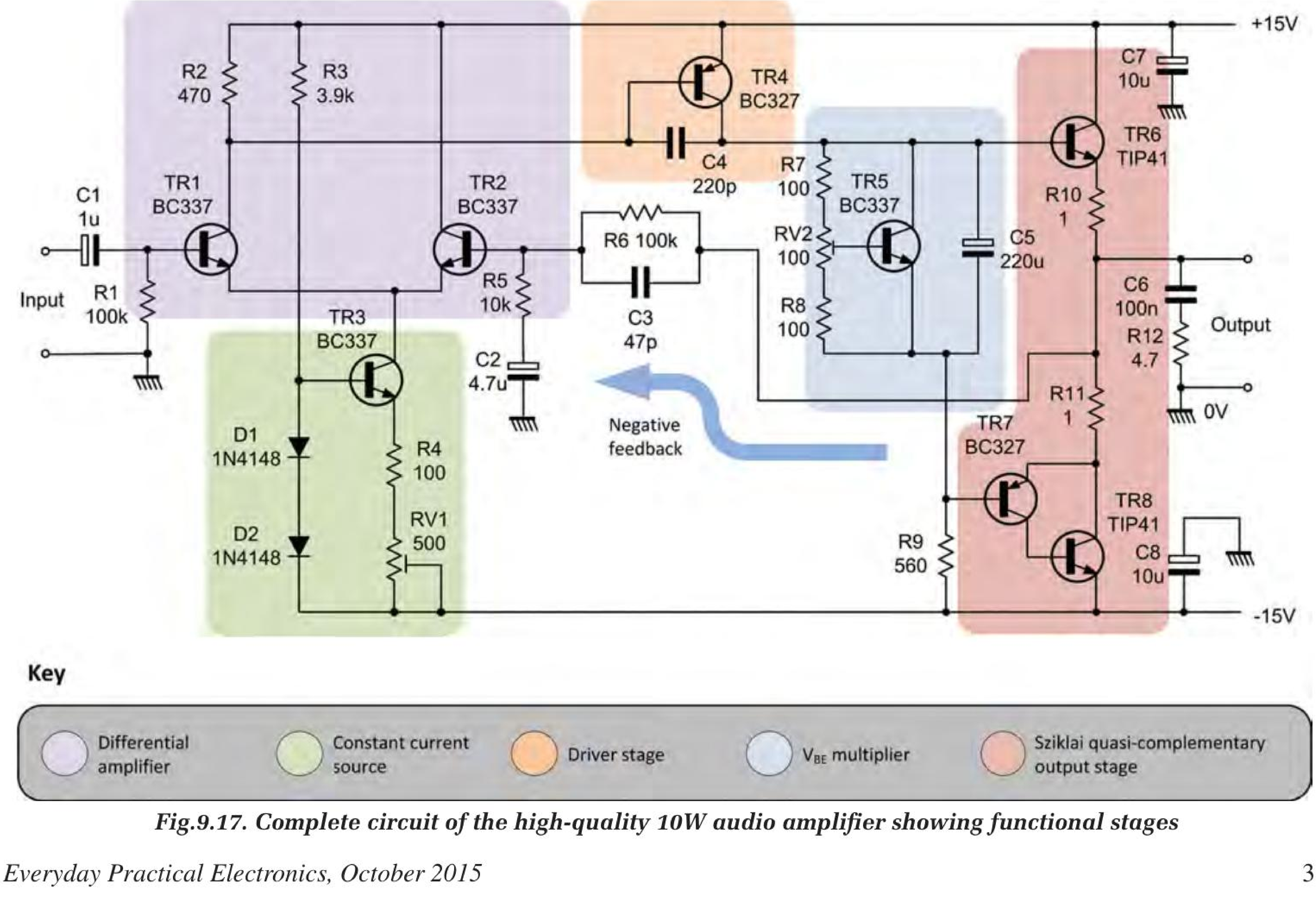


Fig.9.17. Complete circuit of the high-quality 10W audio amplifier showing functional stages

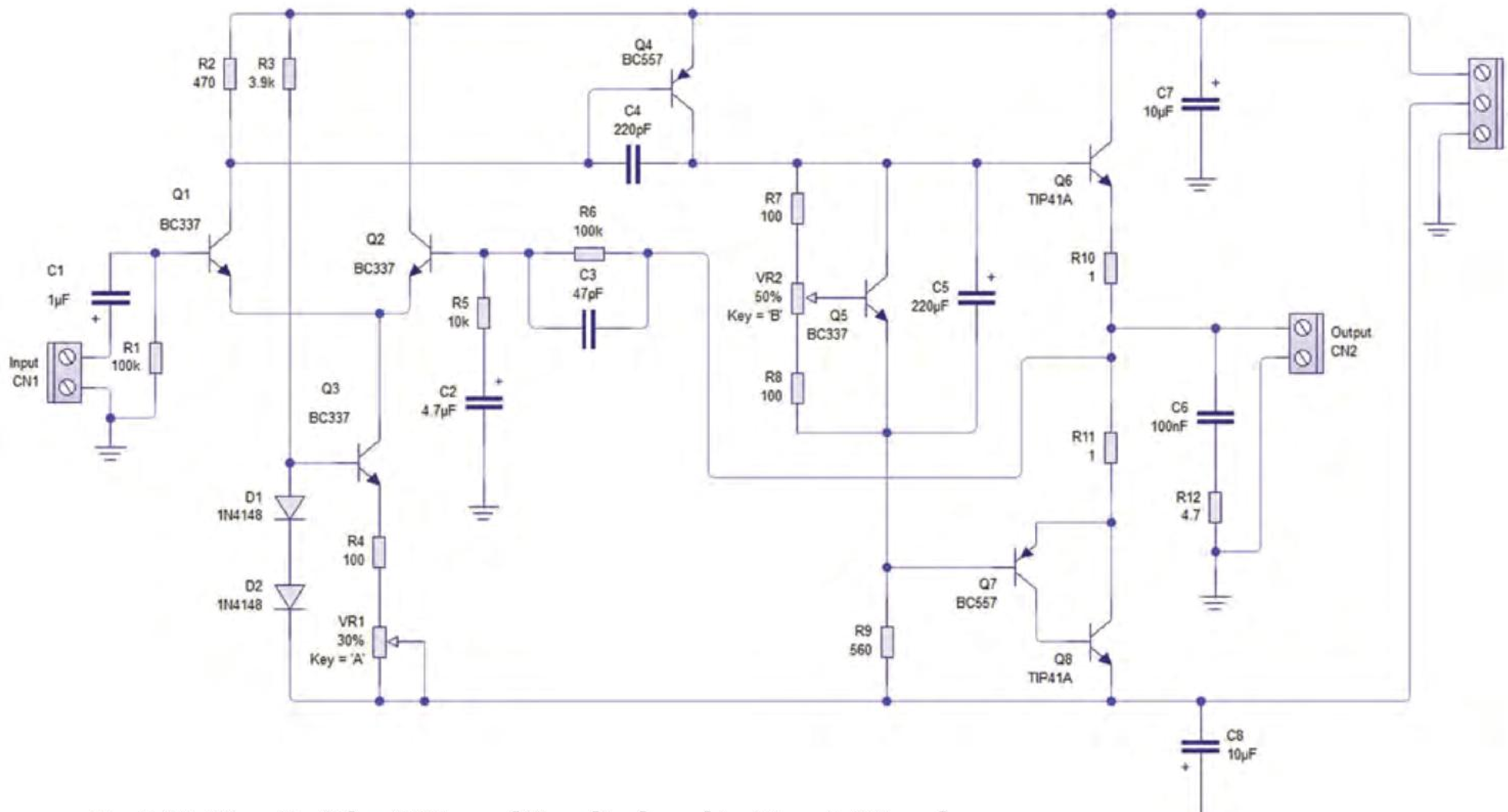

R2
470
R3
3.9k
Q4
BC557
C4
220pF
C7
10µF
+/-15V
CN3
Q1
BC337
Q2
BC337
R6
100k
C3
47pF
R7
100
VR2
50%
Key = 'B'
Q5
BC337
C5
220µF
Q6
TIP41A
R10
1
Output
CN2
C1
1µF
R1
100k
Input
CN1
R5
10k
C2
4.7µF
R8
100
C6
100nF
R11
1
R12
4.7
Q3
BC337
D1
1N4148
D2
1N4148
R4
100
VR1
30%
Key = 'A'
Q7
BC557
R9
560
Q8
TIP41A
C8
10µF

Fig.9.18. Circuit of the 10W amplifier displayed in Circuit Wizard

Semiconductors

4 BC337 (TR1, TR2, TR3, TR5)
2 BC327 (TR4, TR7)
2 TIP41 (TR6, TR8)
2 1N4148 (D1, D2)

Heatsinks

The two power transistors, TR6 and TR8, need to be fitted to individual heatsinks rated at 8.3°C/W or better. These heatsinks should be isolated from the rest of the circuit as this will allow the transistors to be bolted directly to the heatsink surface without the need to use insulating washers and bushes (see *EPE* August 2015). The heatsinks used in the prototype were PP50BPHC units manufactured by ABL. They were supplied pre-drilled for TO220, TO218 and TO247 devices. The heatsinks measure 13 × 35.6 × 50mm and are supported by two solder pins, spaced at 25.4 mm (1 inch). Many other heatsinks are suitable, provided they can be fitted within the available area on the PCB and that they are suitably rated. Under no circumstances should a heatsink rated at worse than 8.5°C/W be used. Please also note that the heatsinks should be of the 'notched' variety with a cut-out so that the three transistor leads can be positioned in line with the two mounting pins when the heatsink is mounted flush with the PCB.

Choice of transistor

We selected low-cost BC337 transistors for use as the low-power NPN stages (TR1, TR2, TR3 and TR5) and BC327

Fig.9.19. PCB component layout

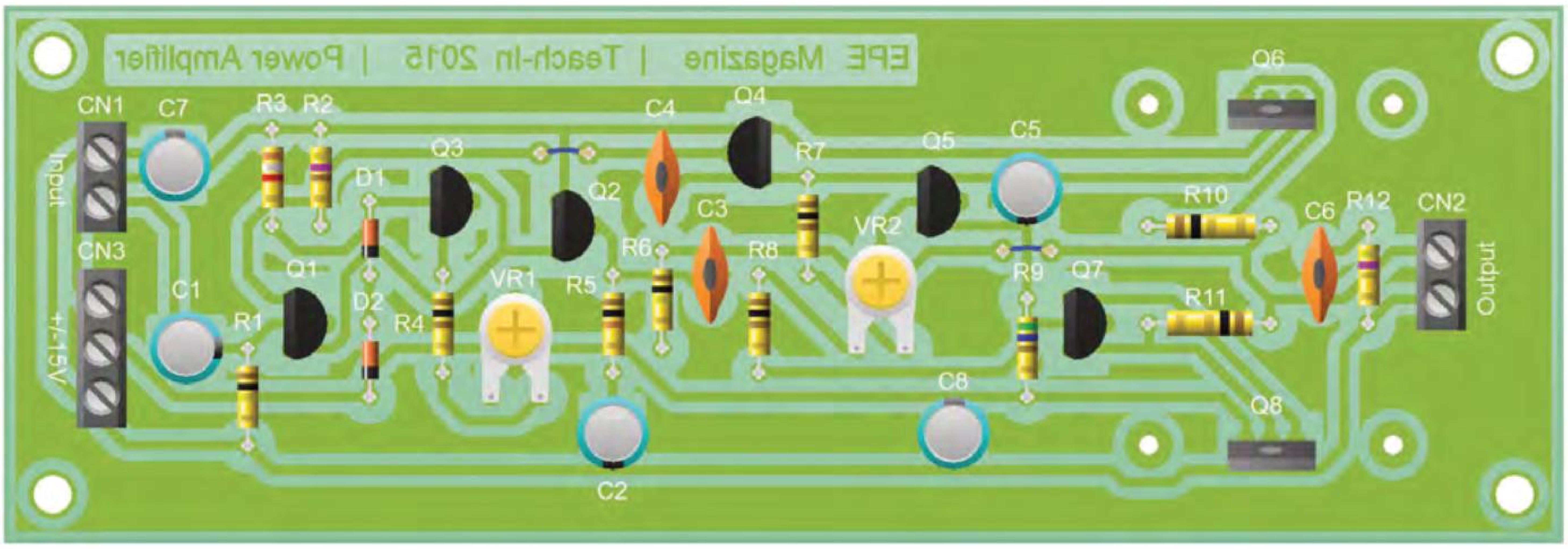
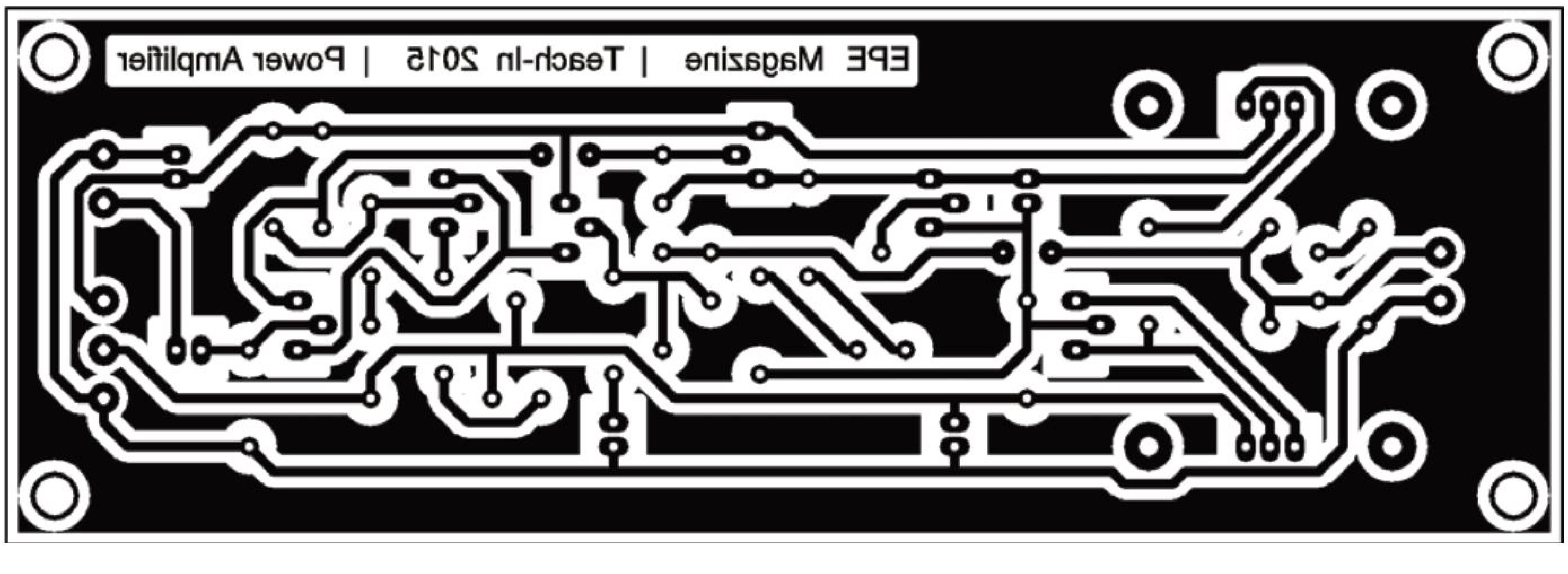

EPE Magazine | Teach-In 2015 | Power Amplifier
CN1
Input
C7
R3
R2
D1
Q3
Q2
C4
Q4
R7
Q5
C5
Q6
R10
C6
R12
CN2
Output
CN3
+/-15V
C1
Q1
R1
D2
R4
VR1
R5
R6
C3
R8
VR2
R9
Q7
R11
C2
C8
Q8

Fig.9.20. PCB track layout


EPE Magazine | Teach-In 2015 | Power Amplifier

devices for use in the PNP stages, TR4 and TR7. These devices are rated for an absolute maximum collector power dissipation of 625mW, which should be de-rated above 25°C by 5mW/°C. Many other devices with current gains in the range 100 to 250 can be substituted (such as BC548 and BC557 devices for the NPN and PNP transistors respectively). Note that the BC548 has a maximum collector power dissipation of 500mW and the quiescent power dissipation for TR3 will be around 400mW in normal operation.

The two output transistors, TR6 and TR8, should have similar current gains (h_{FE}) but, as mentioned last month, devices taken from the same manufacturer's batch will usually be reasonably closely matched. Many other NPN power transistors with similar ratings can be substituted for the specified devices but, as always, it is essential to check on the device pin-out before making any substitution.

Construction

Our prototype PCB was designed to be built into a small enclosure or incorporated into a larger enclosure along with other circuitry. It measures just 163mm × 56mm. In common with all of our *Get Real* projects, the PCB component layout and copper track layout was produced using Circuit Wizard – shown in Fig.9.19 and 9.20 respectively. The printed circuit can be purchased, ready drilled, from the *EPE PCB Service*, code 909. Our finished prototype, ready for testing, is shown in Fig.9.21. Finally, Fig.9.22 shows our pre-prototype version assembled on matrix board. We used this version to carry out performance checks before finalising the design and committing to the final printed circuit layout.

Assembly is straightforward, but we suggest fitting the components to the printed circuit board in the following order: terminal blocks (3), links (2), fixed resistors (12), pre-set resistors (2), diodes (2), non-electrolytic capacitors (5), electrolytic capacitors (5), NPN small-signal transistors (4), PNP small-signal transistors (2), power transistors (2, pre-mounted on two heatsinks). *Note that it is essential to observe the correct polarity of the diodes and electrolytic capacitors and the correct orientation of transistors.*

Setting-up

Before testing it is worth carrying out a careful visual check of the board and components, checking the underside of the PCB for solder bridges between tracks. You will need a digital multi-meter and a fused (or electronically protected) DC power supply with symmetrical outputs of nominally ±15V at 1A (see earlier) connected to CN3, taking care to ensure that the polarity is correct. You will also need a loudspeaker of nominally 4Ω to act as a load, but if this is not available, three parallel-connected 15Ω 5-10W resistors can be used to replace the loudspeaker.

1. Set RV1 to mid-position and RV2 fully clockwise.
2. Connect the digital multi-meter (on the 2V DC range) to read the DC voltage appearing across the output at CN2 (the negative meter lead should be taken to the terminal that appears closest to TR8). Switch the power supply 'on' and carefully adjust RV1 for a reading of less than ±10mV.
3. Re-connect the digital multi-meter (on the 2V DC range) so that it reads the DC voltage appearing across R10 (the negative meter lead should be taken to the junction of R10 and R11 near C6 on the printed circuit board). Switch the power supply on and adjust RV1 for a reading of between 25mV and 35mV (corresponding to a standing bias current of between 25mA and 35mA).
4. Reconnect the digital multi-meter (on the 2V DC range) to once again read the DC voltage appearing across the output at CN2 (the negative meter lead should be taken to the terminal that appears closest to TR8). Switch the power supply 'on' and readjust RV1 for a reading of less than ±10mV. Note that there is some interaction between the two pre-set adjustments and so further adjustment might be necessary.
5. Wait for about 10 minutes in order to allow the amplifier to reach normal operating temperature and then repeat steps 2 to 4. The aim of the adjustment procedure is to ensure that, with no signal applied, the output voltage is less than ±10mV and that the standing (quiescent) bias current in the output stage is approximately 30mA.

Finally, our measured test voltages are shown in Table 9.2.

Table 9.2 Test voltages

Device	Emitter	Base	Collector
TR1	–1.0V	–0.4V	+16.7V
TR2	–1.0V	–0.4V	+17.1V
TR3	–16.2V	–15.6V	–1.0V
TR4	+17.1V	+16.7V	+0.6V
TR5	–0.6V	0V	+0.6V
TR6	+0.03V	+0.6V	+17.1V
TR7	–0.03V	–0.6V	–16.45V
TR8	–17.1V	–16.45V	–0.03V

Next month

In the final instalment of *Teach-In 2015*, *Discover* will examine practical aspects relating to measurement, adjustment and fault-finding in power amplifiers. *Knowledge Base* will look at power supplies, and *Get Real* will explain the procedure for measuring the performance of our high-quality 10W amplifier and we reveal whether or not it met or exceeded our original design specification.

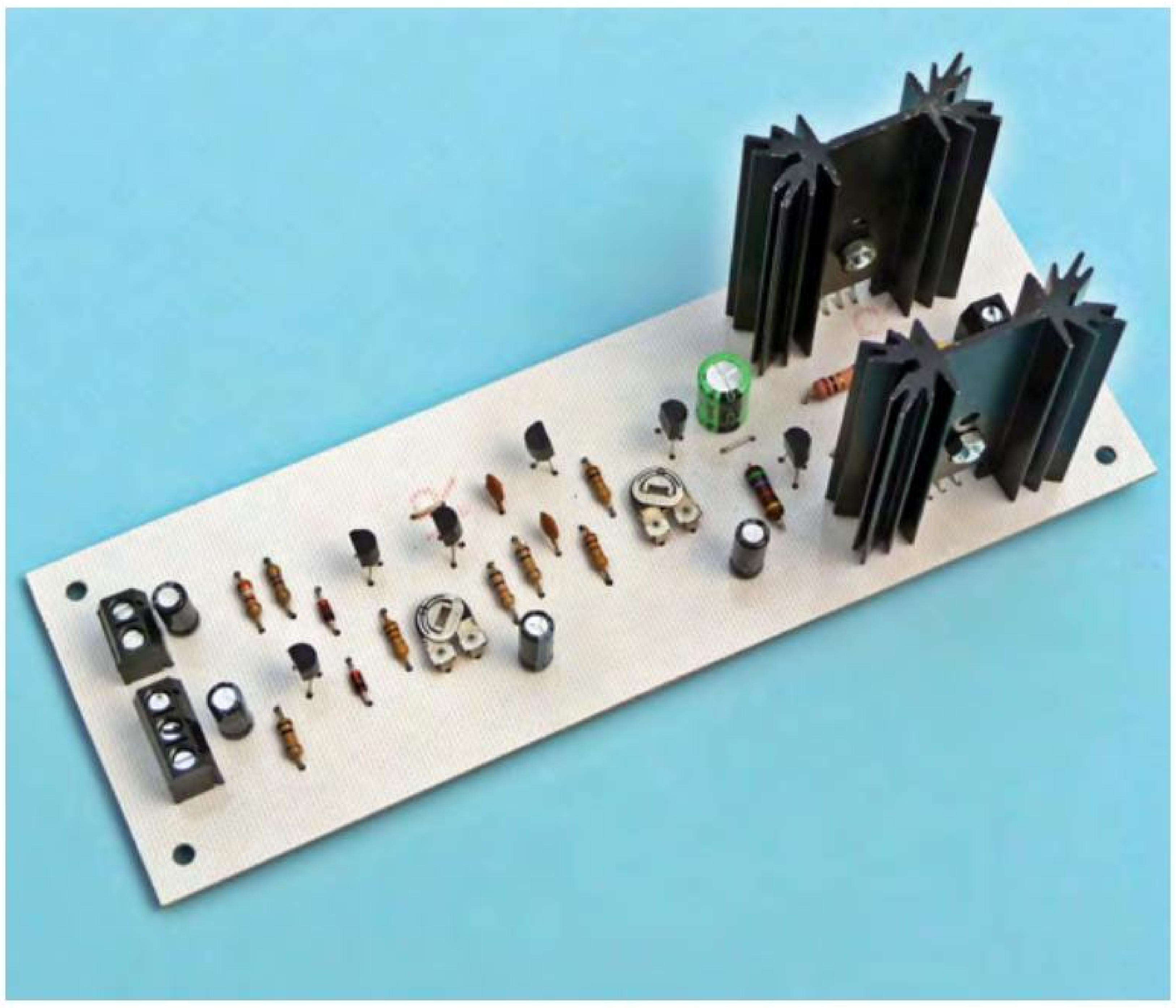


Fig.9.21. Completed 10W amplifier ready for testing

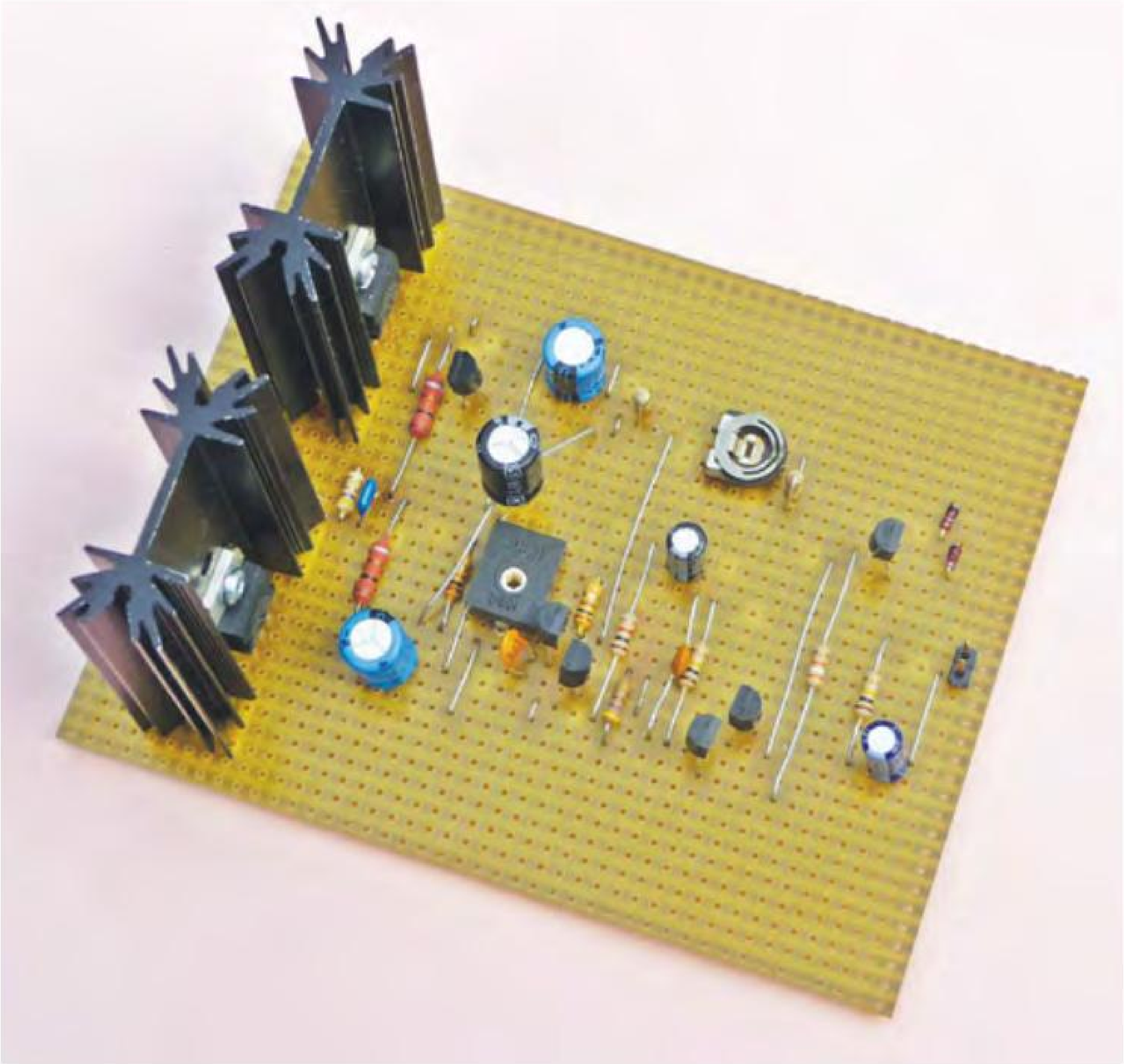


Fig.9.22 The pre-prototype 10W amplifier built on low-cost matrix board

NET WORK

by Alan Winstanley

When content is king

BRITISH Telecommunications (BT) claims to be the world's oldest communications company, tracing its heritage all the way back to 1846. At a local conference some 15 years ago, BT sought to squeeze more from its broadband by pleading with developers and web designers to take full advantage of the then-new ADSL by providing more multimedia content for end users; the network speed was there, but online content that could fully utilise that speed had yet to arrive. Today, with high-speed VDSL now on offer over its phone lines, BT wants a greater stake in building today's 'information age' and it has the financial firepower to do it. In 2013, BT paid nearly £900 million ($1.4 billion) for the rights to stream three years' worth of live Champions League and Europa League football. BT's website rubs it in when bragging about its online programming, reminding us that *'BT Sport is the only place to watch all the UEFA Champions League live. These matches are no longer shown on ITV or Sky Sports.'* Not resting on its laurels, the telecoms titan recently announced a new BT TV Ultra HD service as well, that they claim includes Europe's first live sports Ultra HD (4K) channel. So now you can watch a football being kicked around with four times the definition of HD.

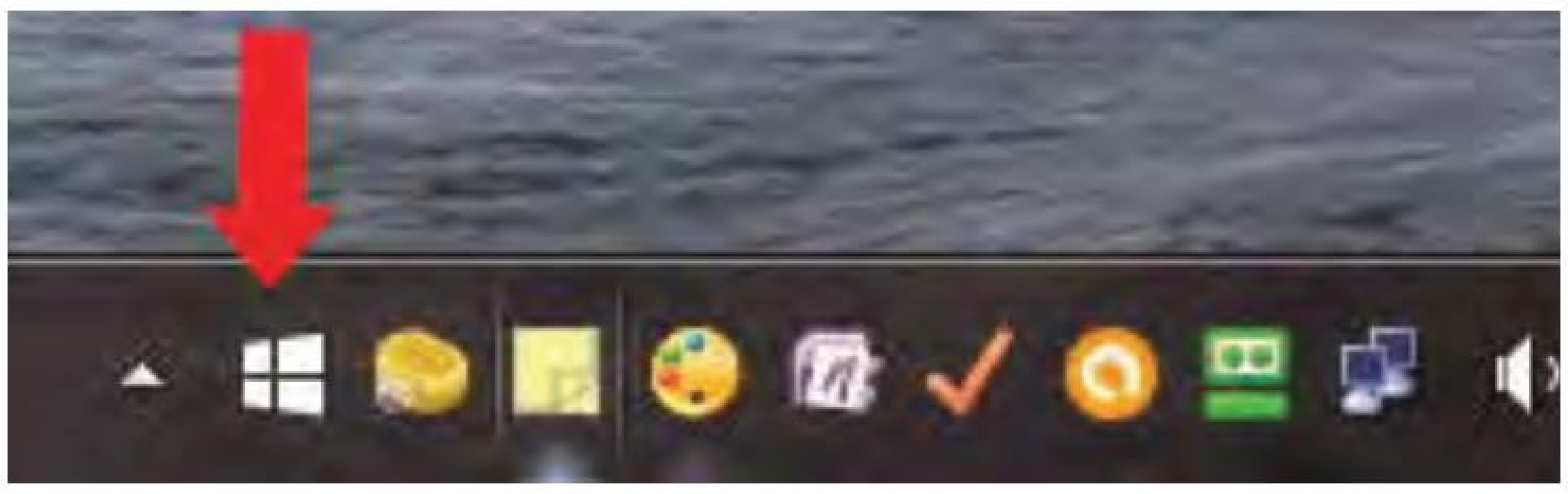
If BT wants to captivate us with online content then it faces plenty of competition. Amazon continues its very aggressive bid to grab more market share with the news that the former BBC TV 'Top Gear' team has been signed up to produce three series exclusively for Amazon Prime. The hugely popular Top Gear motoring programme had been a BBC TV staple and the brand was sold very successfully worldwide, but the series fell apart after presenter Jeremy Clarkson took a swing at a TV producer after a tough day's filming. The orphaned programme had been looking for a new home and rumours were rife about switching to a streaming service. In an extremely deft move, a new series will premier on Amazon Prime in 2016, and its success is assured. Amazon Prime in the UK currently costs £79 a year and it bundles TV and movies with a one-day delivery service on much of its merchandise. Amazon has also introduced Prime Music, offering a million songs at no extra charge for its Prime members.

amazon Prime
AMAZON ORIGINAL

BBC's former Top Gear programme moves to a new format on Amazon Prime in 2016. Picture credit: Amazon Prime Video

Taking Ten

As the joke goes, you can spend more time on Netflix looking for movies than actually watching them, and I have rented my share of duff films that were a waste of money. After an evening of sifting through endless choices of online movies, Microsoft Studios' *Halo 4: Forward Unto Dawn* finally burst onto my TV screen. Featuring the characters first seen in the Halo Xbox video games, the movie gave the nod to an interactive artificial intelligence called Cortana that conversed with our armour-plated hero. The holographic Cortana first appeared in Halo on Xbox in 2001, but it's also the name for Microsoft's digital assistant found in Windows 10 and Windows Phone.



The Get Windows 10 app appears in the system tray of Windows 7 (SP1) or higher. Check Windows Update if it is missing

Microsoft's latest release of its ubiquitous Windows desktop operating system formally bowed in on 29 July, and is being made available as a free upgrade until July 2016. Qualifying users of Windows 7 (SP1) and Windows 8.1 might have already seen a teasing new Windows icon – the 'Get Windows 10 app' – appearing in the system tray as a symbol heralding better things to come. The 'Check my PC' option in the app runs a gamut of compatibility checks to reveal whether or not the PC and its software is upgradeable to Windows 10. Clicking the Get Windows feature initiates your Windows 10 reservation and a mighty 3GB upgrade will be downloaded onto your PC as a matter of course. Windows 10 is then yours to enjoy for free.

If the tantalising Get Windows 10 app has not appeared on your PC then Windows Update is probably disabled or Microsoft Windows Update has already decided that your PC cannot run it. The icon should still have appeared automatically on the due date in July, primarily to help users figure out what was needed to make their system compliant, if they chose to do so. The author duly ran Windows Update manually and the Windows 10 icon put in an appearance after rebooting. On the big day itself, a new popup report confirmed that my PC was fully compatible with Windows

10. Like many users, though, I remain perfectly content with Windows 7, and feel no need to upgrade straight away.

Among other benefits, Windows 10 brings back the Start button and menu, the user interface (UI) first introduced 20 years ago in Windows 95 and foolishly omitted in the unloved Windows 8.0. Windows 10 also introduces Microsoft Edge, their new, leaner, faster web browser that promises to be everything that Internet Explorer is not (fast, secure and reliable). It has a two-dimensional characterless styling first mooted by Google Chrome, but Microsoft Edge integrates with Cortana to provide voice interaction in the same way that Google, Apple's Siri and Amazon's Alexa offer voice control for their users. Users can also 'write' on web pages in Edge and share notes with others.

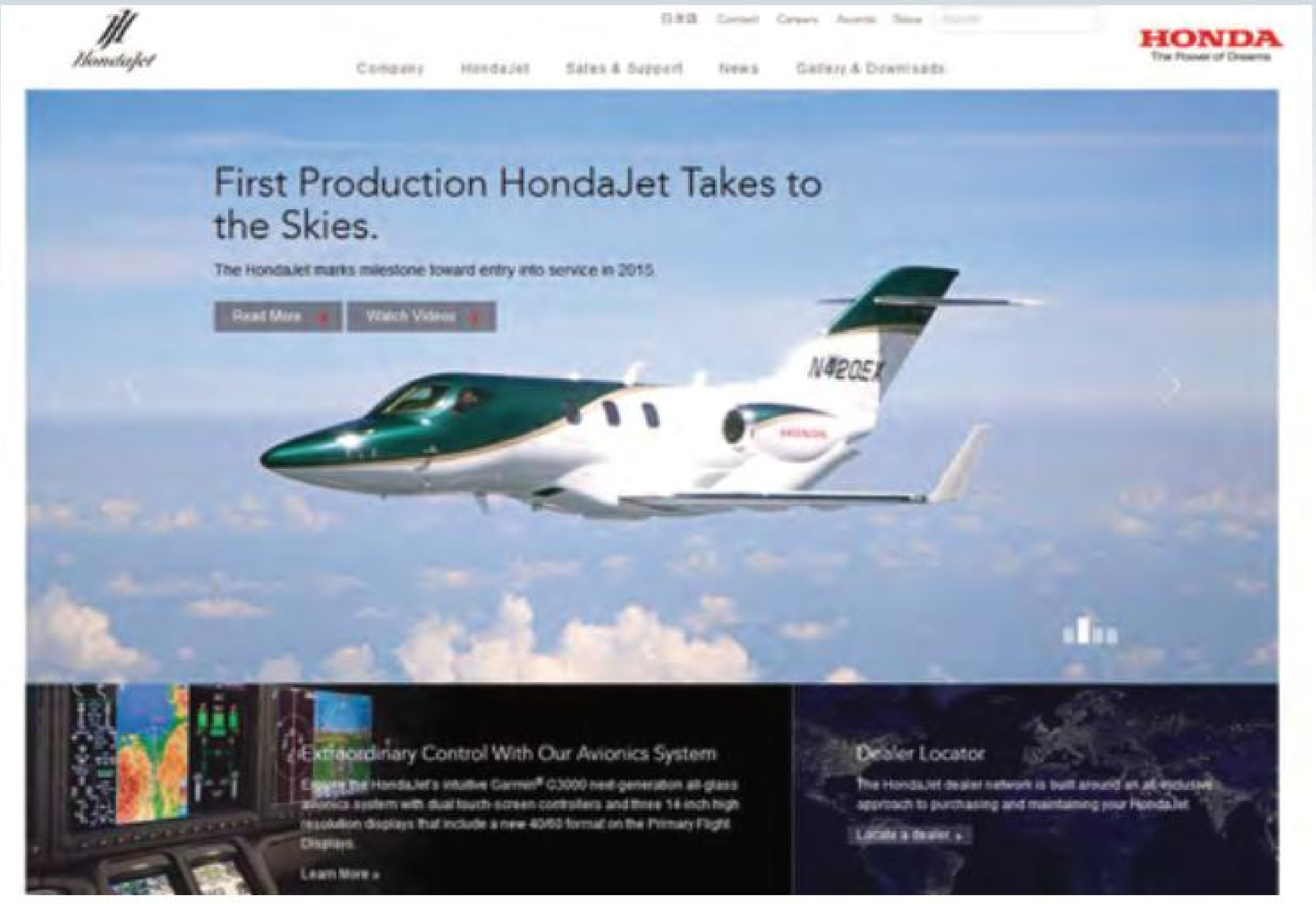
There is no doubt that the stakes could not be higher for Microsoft's latest operating system, and downloading the free upgrade to Windows 10 within the next twelve months will be the best choice for PC users wanting to move with the times and keep their system as secure as possible. Maybe users may not be ready to upgrade just yet, but Microsoft has graciously allowed a year to think about it. Future upgrades of Windows 10 will be incremental: there will be no new release and instead Microsoft will enhance Windows in the same way that many operating systems and apps update themselves on a rolling basis.

Flat Earth society

Windows users are not alone when it comes to the need to update operating systems. Owners of Android handsets and tablets bear witness to the drudgery of apps, sometimes several dozen of them at a time, being patched to the latest versions. To add insult to injury, the author's Android phone updated itself one evening, which resulted in its Do Not Disturb (DND) function disappearing. The DND feature stops the smartphone from sounding at inconvenient times. Consequently, the newly-updated phone woke the writer at 3am with the sound of new email arriving, and 4am saw him stuffing it under a pillow and googling to find out how to silence the wretched thing – now running Android Lollipop – when it's supposed to stay silent overnight. For some reason, following the Android update, DND had become 'Interruptions', accessible via the phone's volume controls. The option to silence it automatically on Sundays had scrolled off the bottom of the Android screen, tucked out of sight. Hence the phone would wake the writer on Sunday nights, beeping incessantly with new email until eventually the hidden 'Sunday' tick-box was found on the calendar menu, and weekend bliss was finally restored.

This fashion for flat, bland styling is creeping into UI design on websites, PCs and touchscreens everywhere, and a counter-intuitive and featureless landscape of coloured tiles and oversized text with few usage cues is all too commonplace these days. Traditional Windows-type sliders and controls are being usurped by flat, touchscreen-type navigation, begging you to scroll up or down with your thumb. This lack of a recognisable slider knob is typical of the over-simplified 2D user-interface beloved of the current generation of website and touchscreen designers. Furthermore, the need to re-learn a user interface following an Android update and get to grips with any changes is becoming a very annoying and time-consuming aspect of device ownership.

In recent years – until the touchscreen came along – the vast increases in computer graphics and monitor performance had seen more attractive 3D graphics, alpha blending and transparency, drop shadows and high-resolution icons (eg the Windows Aero interface) that made computer use both intuitive and attractive. The information age has now gone 'flat earth' and users are presented with expanses of flat colour tiles and often poorly-contrasting icons. Maybe you will spot the odd 'X' to close a window if you're lucky, but operating systems, software and apps are all heading the same way. Hand in hand with this is the fashion for the so-called 'hero image' – a term borrowed from the print design industry for a very large graphic or background image that dominates a page. Full-screen hero images often have no menu or navigation at all, but are only feasible due to improved Internet bandwidth and larger computer screens.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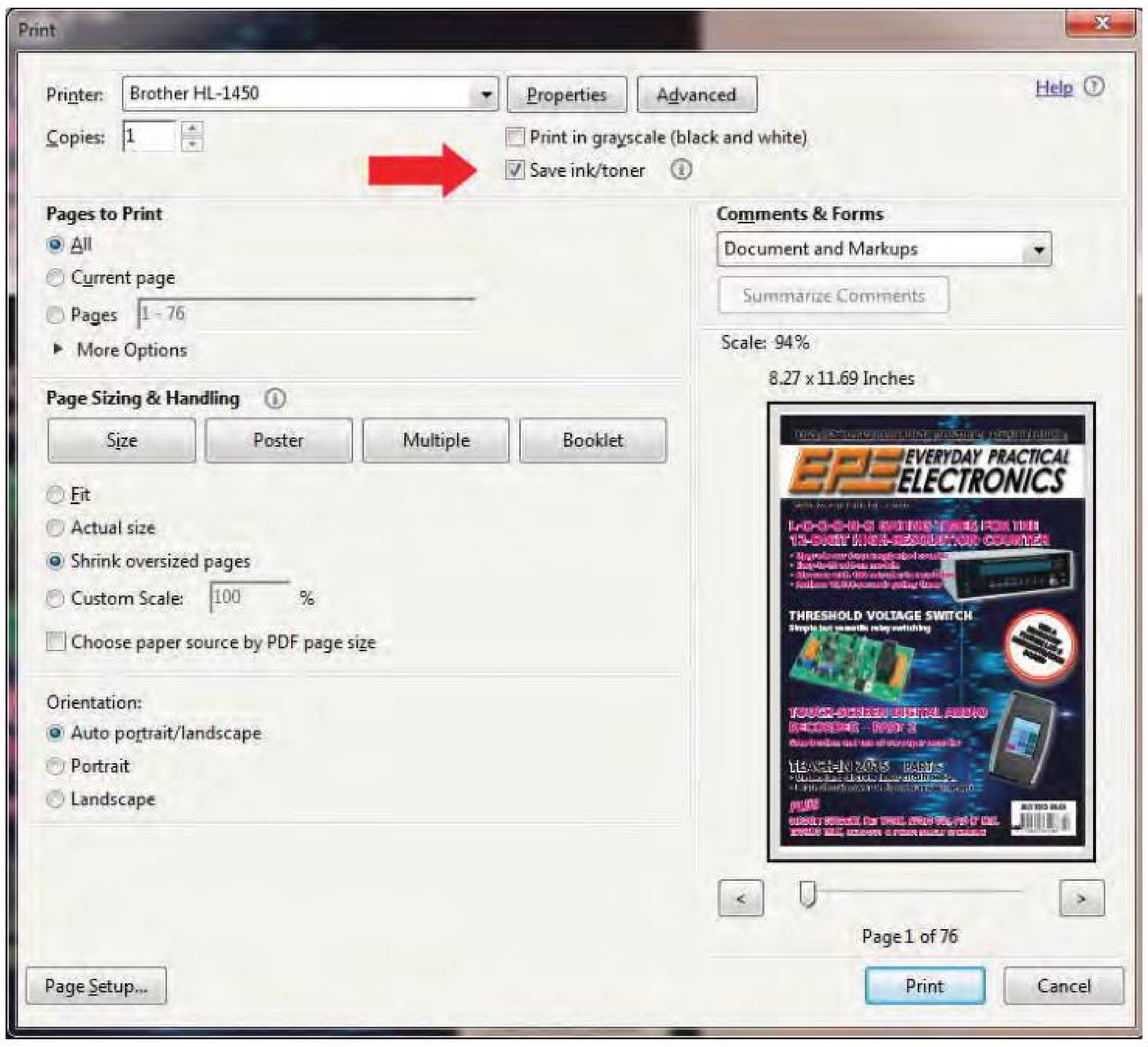
A large 'hero image' dominates a web page and is an increasing trend in web design

Out in a Flash

Regular *Net Work* readers will know how I constantly nag about the need to keep software updated, mainly to defend a system against vulnerabilities that could be exploited by hackers. A 'zero day' vulnerability is one that is currently unpatched by vendors, which leaves the product susceptible to attack. Most software houses release updates fairly swiftly and users would have to be quite unlucky to suffer any problem directly, but heading my list of likely targets are Adobe Reader and Adobe Flash Player. They offer many opportunities for hackers to exploit, either by opening an infected document on your PC or by visiting a website that exploits weaknesses in your software.

The latest Adobe Acrobat Reader DC (Document Cloud) is best known as a standalone PDF viewer that also has a web browser plugin. It includes some useful features, such as Toner/Ink Saving, which saves up to 15% by economising on the amount of ink or toner printed onto paper. (The Ryman Eco typeface from **www.ryman.co.uk/ryman-eco/** can be used in your own documents to reduce by 33% the volumes of ink or toner needed to print them.) The free Adobe Reader DC also tantalises users with useful tools such as rubber-stamping or overprinting PDFs, but some tools such as Convert (files) to PDF head off to an Adobe subscription sign-up webpage costing up to £100 a year!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○ Portrait
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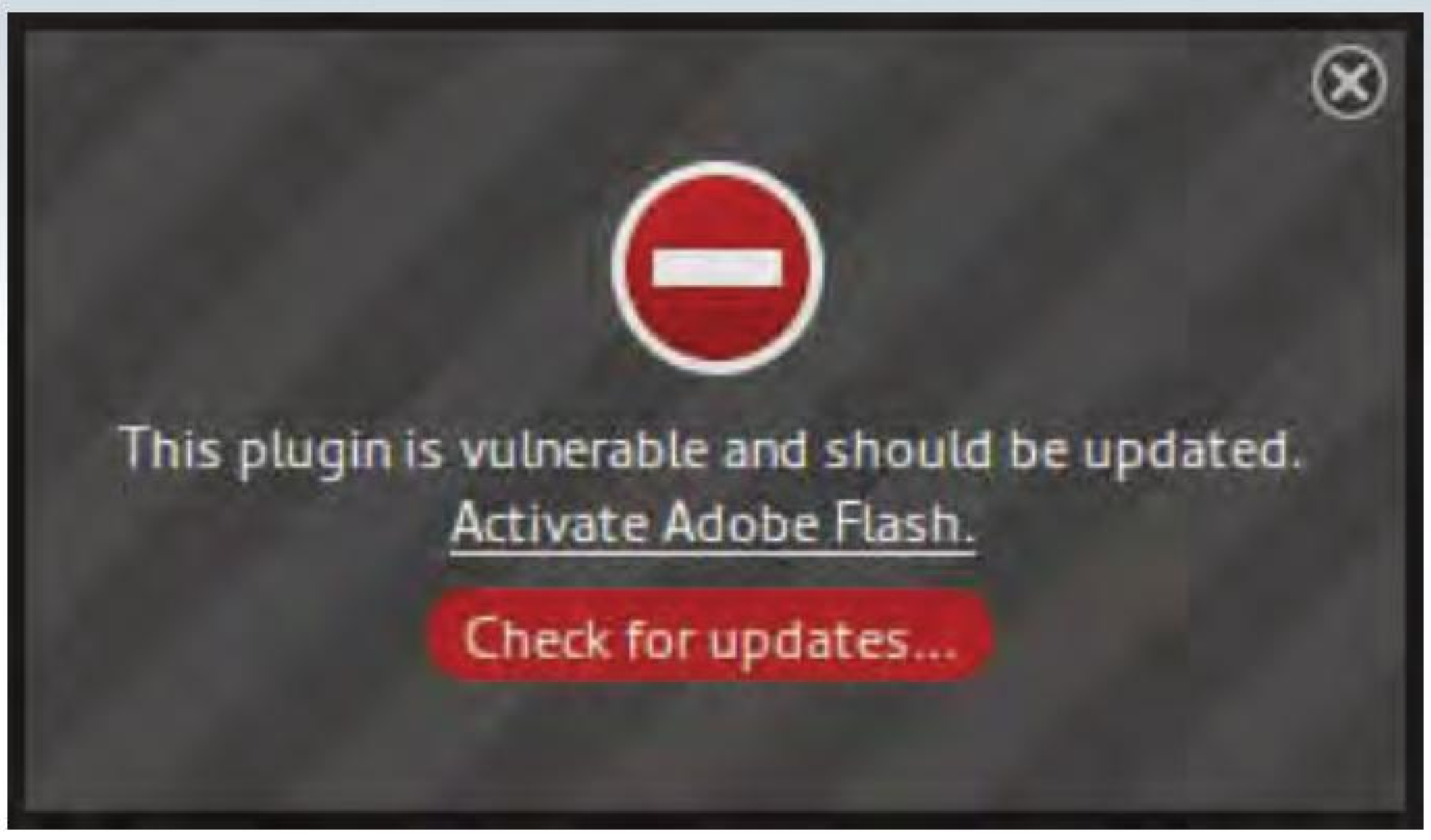
Comments & Forms
Document and Markups

Scale: 94%
8.27 x 11.69 Inches
Page 1 of 76

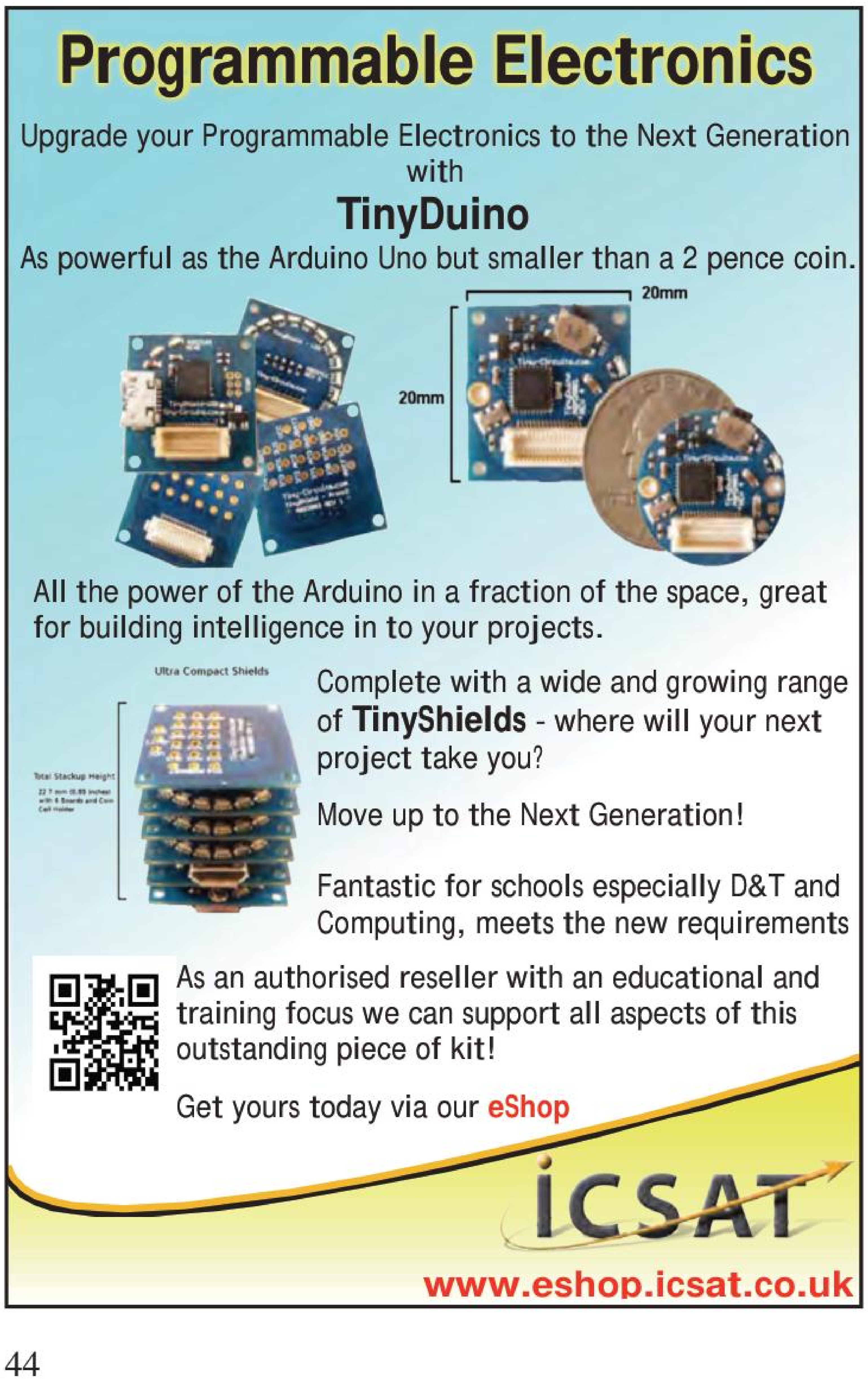
Adobe Acrobat Reader DC has a printing economiser function that it claims saves up to 15% of ink or toner when printing PDFs


This plugin is vulnerable and should be updated.
Activate Adobe Flash.

Firefox now blocks by default the Flash plugin unless you activate it

So-called 'Flash' is the technology that delivers motion graphics, video and audio over the web. In recent weeks both Firefox and Google web browsers decided to block Flash altogether due to perceived security risks. This is now resulting in web pages failing to load inline content such as Google adverts and banners or multimedia presentations. Users can manually allow Flash to run, but this can be such a nuisance that web surfers prefer to simply ignore the Flash plugin warning and not bother activating Flash at all. The most advanced media websites are gradually adopting the latest web standards – HTML5 – which avoids the need for Flash altogether. For now though, users may have to make their own judgments about whether to allow Flash to run on a particular web page, and you can be sure that Firefox and Google Chrome will put obstacles in the way of displaying these multimedia elements.

That's all for this month's *Net Work*. You can contact the writer at **alan@epemag.demon.co.uk**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By Robert Penfold

INTERFACE

Raspberry Pi PWM

USING the Raspberry Pi to produce a controllable output voltage with the aid of a digital-to-analogue converter (DAC) has been covered in previous *Interface* articles. This method is the best approach for many applications, but is less than ideal for others. The alternative approach of using pulse-width modulation (PWM) is preferable for some applications, such as controlling the speed of DC electric motors, and the brightness of filament bulbs or LEDs. PWM often requires simpler and cheaper hardware, and in power-control applications it is more efficient.

PWM basics

PWM does not provide a variable DC output voltage, but instead delivers a pulse signal that has a variable duty cycle. It is the average output voltage that is controlled, although with suitable filtering a DC voltage equal to the average output potential can be provided. Fig.1 shows some example waveforms that help to explain the way in which PWM operates. In the top waveform there is a 1:1 mark-space ratio, and the output is switched off for 50 percent of the time. The average output potential is therefore equal to half the ‘on’ voltage. In the middle waveform the mark-space ratio is 4:1, and the average output potential is therefore 4/5 or 80 percent of the ‘on’ voltage. The lower waveform is the inverse of this, with a 1:4 mark-space ratio, and an average output potential equal to 1/5 or 20 percent of the ‘on’ voltage. Most PWM controllers use a fixed frequency and a variable duty cycle, as in the example waveforms, but much the same effect can be obtained using a fixed pulse width and a variable frequency.

Efficiency is high with a PWM controller because the output transistor operates as a switch, and it provides either the full output potential or no output voltage at all. In theory, there is no power wasted in the control circuit. The switch passes a high current when it is closed, but the voltage drop is zero, giving zero power loss. No output current at all flows when the switch is open, and the output power is therefore zero. Of course, theoretical perfection is one thing and real-world electronics is another. Although there is a totally insignificant current flow when a switching transistor is turned off, there is a significant voltage drop when a transistor is switched on, especially when high currents are involved. PWM is still much more efficient than using some form of variable resistance, where the power wasted in the controlling element will sometimes exceed that supplied to the load.

Of course, in its basic form with no output filtering, pulse control is not suitable for all applications. Where it is suitable, due care has to be taken when selecting the pulse frequency. Using a very low frequency is unlikely to be successful. LEDs and filament bulbs would be seen to flash on and off rather than dim. Similarly, a DC electric motor would operate in a ‘fits and starts’ fashion rather than operating at a set speed. High pulse frequencies could give problems with some types of load, for example, inductive types. There could also be problems with the generation of strong radio frequency interference (RFI). A frequency in the region of 50 to 200Hz is suitable for most practical applications of pulse control.

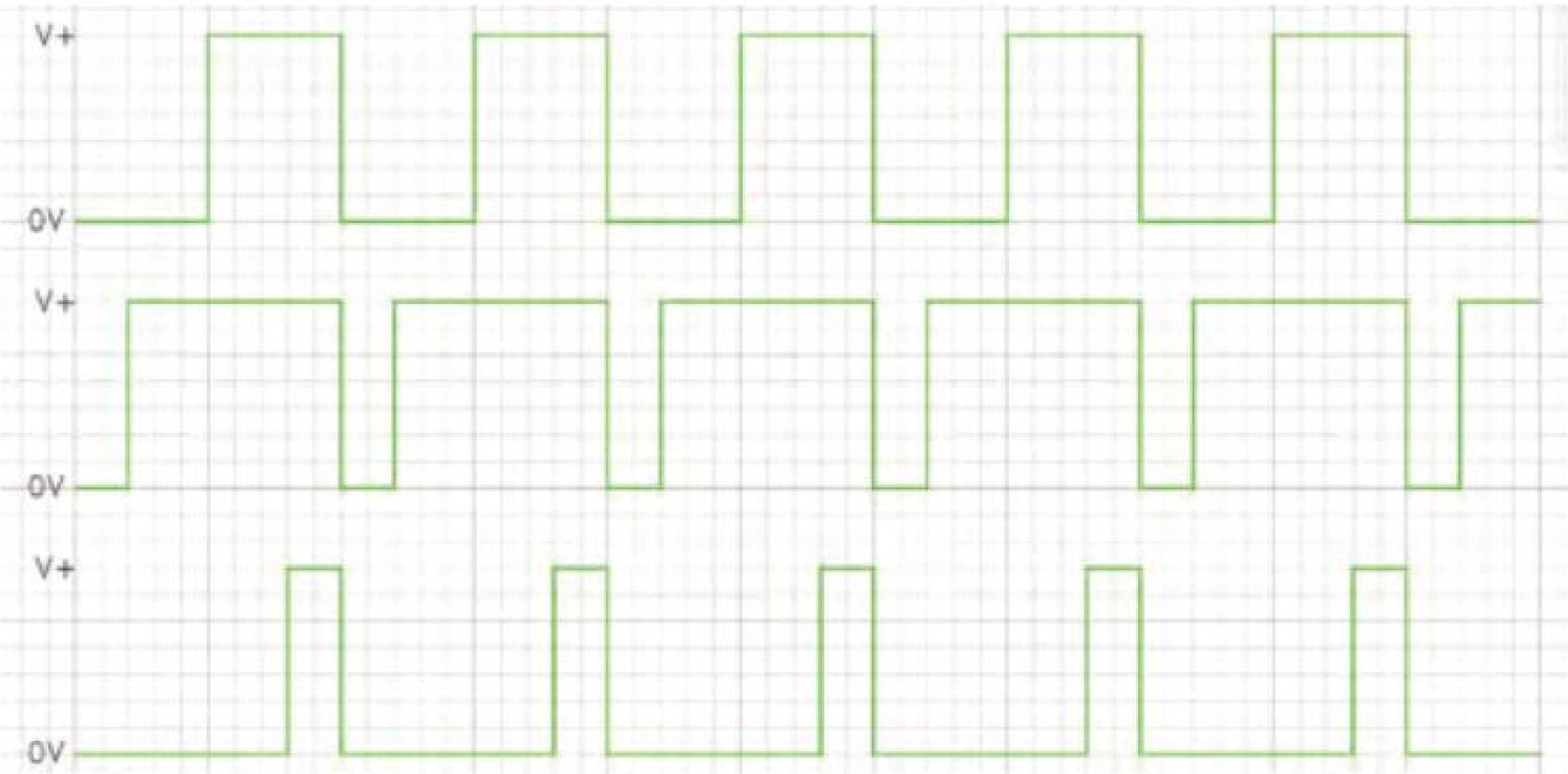


Fig.1. With PWM it is the average output voltage of the signal that is of importance. In these examples, from top to bottom, the average output potentials are 50, 80 and 20 percent of the V+ level

Software PWM

PWM can be achieved using hardware or software. The hardware method has been covered in previous *Interface* articles, and one method is to output values to a digital-to-analogue converter (DAC). The output voltage of the converter is fed to a circuit that is comprised of a triangular waveform generator and a voltage comparator. Much simpler and cheaper hardware is required with the software method, because all the hard work is done in the program. This generates the required waveforms on a digital output of the computer, and the hardware simply has to provide voltage amplification and buffering so that the load can be driven properly.

The Raspberry Pi has PWM software available via Python and the `RPi.GPIO` module. This enables a PWM controller to be implemented very easily using minimal hardware. The software is very simple as well, because the wave generation is handled by the `RPi.GPIO` module, and the control program simply has to instigate the required changes in the duty cycle via the appropriate Python program instruction. Any of the GPIO input/output pins can be used as a PWM output, and it is possible to have two or more PWM outputs operating simultaneously. Any line used as a PWM output must first be set to the output mode in the normal way.

One problem with software PWM is that the accuracy might be impaired by the processor servicing interrupts and handling other tasks that keep the computer running properly. In practice, Raspberry Pi software PWM works well, but it is probably best to keep the operating frequency quite low. Higher frequencies involve shorter pulse durations that are more vulnerable to brief processing interruptions. Also, there is a limit to the minimum pulse duration that can be provided, and the maximum available resolution of the system is therefore better at low output frequencies.

PWM instructions

There are several instructions available for use with a PWM output:

`p = GPIO.PWM(channel, frequency)`

Here, `p` is the name given to the PWM output, and it is effectively the variable used to control it, and `channel` is the number of the GPIO line that is being used to output the PWM signal. The frequency is specified in hertz (Hz).

`p.start(duty cycle)`

This instruction starts the PWM output signal and sets the initial duty cycle. The value used here is from 0 (always low) to 100 (always high), and the value does not have to be an integer. This value is setting the percentage of the time that the PWM output is in the high state.

`p.ChangeDutyCycle(duty cycle)`
Once the PWM output has started operating, this instruction is used to alter the duty cycle. It should be noted that the capitalised letters in the instruction are required. Using (say) `changedutycycle` will not work.

`p.ChangeFrequency(frequency)`
In a normal PWM application it is not necessary to change the initial operating frequency, but it can be achieved using this instruction. Again, the capitalisation of some letters is required for the instruction to work.

`p.stop()`
The PWM signal should be terminated using this instruction prior to the program ending. The `GPIO.cleanup()` instruction should still be used at the end of the program in the usual way.

Software

The simple Python program of Listing 1 can be used to test the PWM instructions. The first part of the program imports the `RPi.GPIO` and time modules. The latter is needed because the program uses the sleep function of the time module to provide delays. The main section of the program sets pin 26 of the GPIO port as an output, designates it as a PWM output, sets the initial operating frequency at 100Hz, and then starts the output signal with a duty cycle of 50 percent. After an eight-second delay, the output frequency is increased to 200Hz, and after a further eight-second delay, the duty cycle is changed to ten percent. The PWM signal is switched off after a further eight-second delay, and the program then terminates.

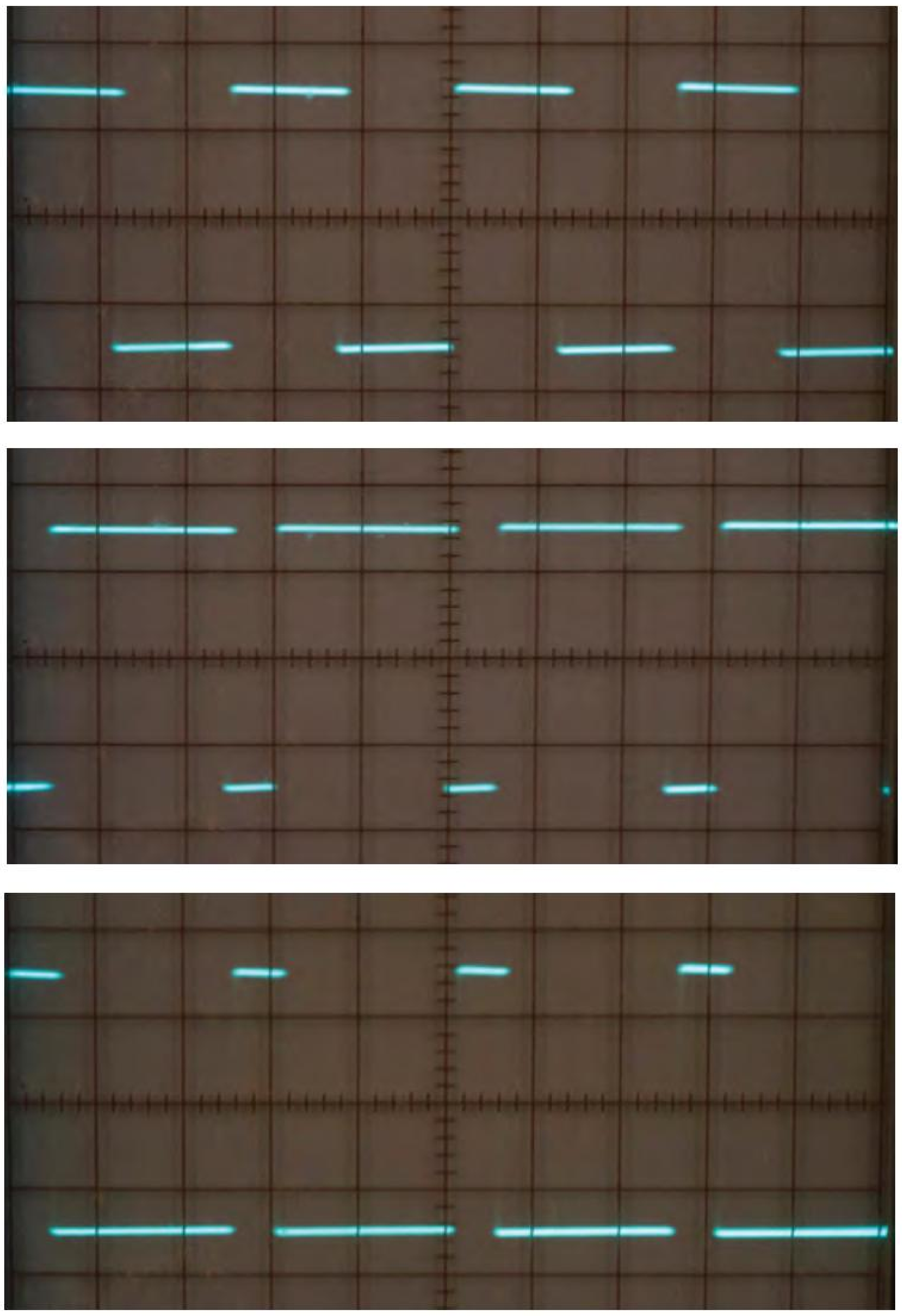


Fig.2. Output waveforms from the Raspberry Pi with duty cycle values of 50 (a), 80 (b) and 20 (c), at an output frequency of 200Hz. Values of 0 and 100 respectively set the output continuously low or high

Listing 1

```
 import RPi.GPIO as GPIO
 import time

 GPIO.setmode(GPIO.BOARD)
 GPIO.setup(26,GPIO.OUT)
 p=GPIO.PWM(26,100)
 p.start(50)
 time.sleep(8)
 p.ChangeFrequency(200)
 time.sleep(8)
 p.ChangeDutyCycle(10)
 time.sleep(8)
 p.stop
 GPIO.cleanup()


 Listing 2
 import RPi.GPIO as GPIO
 import time

 GPIO.setmode(GPIO.BOARD)
 GPIO.setup(18,    GPIO.IN,     pull_up_
down=GPIO.PUD_DOWN)
 GPIO.setup(22,    GPIO.IN,     pull_up_
down=GPIO.PUD_DOWN)
 GPIO.setup(24,    GPIO.IN,     pull_up_
down=GPIO.PUD_DOWN)
 GPIO.setup(26,GPIO.OUT)
 p=GPIO.PWM(26,100)
 p.start(0)
 dutcyc = 0
 x = 1
 print ("Started")

 while (x == 1):
     if GPIO.input(18):
         x = 2
     if GPIO.input(22):
         dutcyc = dutcyc + 1
     if dutcyc == 101:
         dutcyc = 100
     if GPIO.input(24):
         dutcyc = dutcyc - 1
     if dutcyc == -1:
         dutcyc = 0
     p.ChangeDutyCycle(dutcyc)
     time.sleep(0.1)

 p.stop
 print ("finished")
 GPIO.cleanup()
```

Ideally, an oscilloscope is needed in order to check the output signal properly. Figs.2a to 2c show the output waveforms obtained with duty cycle values of 50, 80, and 20 respectively. There are simpler ways of checking the output signal. Connecting the output signal to a crystal earphone or the input of an amplifier/speaker will enable the changes in the signal to be heard. An analogue multimeter set to a low DC voltage range should show a reading of about 1.65V initially, or in other words, about half the 3.3V high logic level of the output. The reading should drop to about 0.33V when the signal switches to a duty cycle of ten percent. This will also work with most digital multimeters, but some will not respond to the average voltage of a pulsed signal.

Another way is to use the simple LED driver circuit of Fig.3. This uses a common-emitter switching transistor Tr1

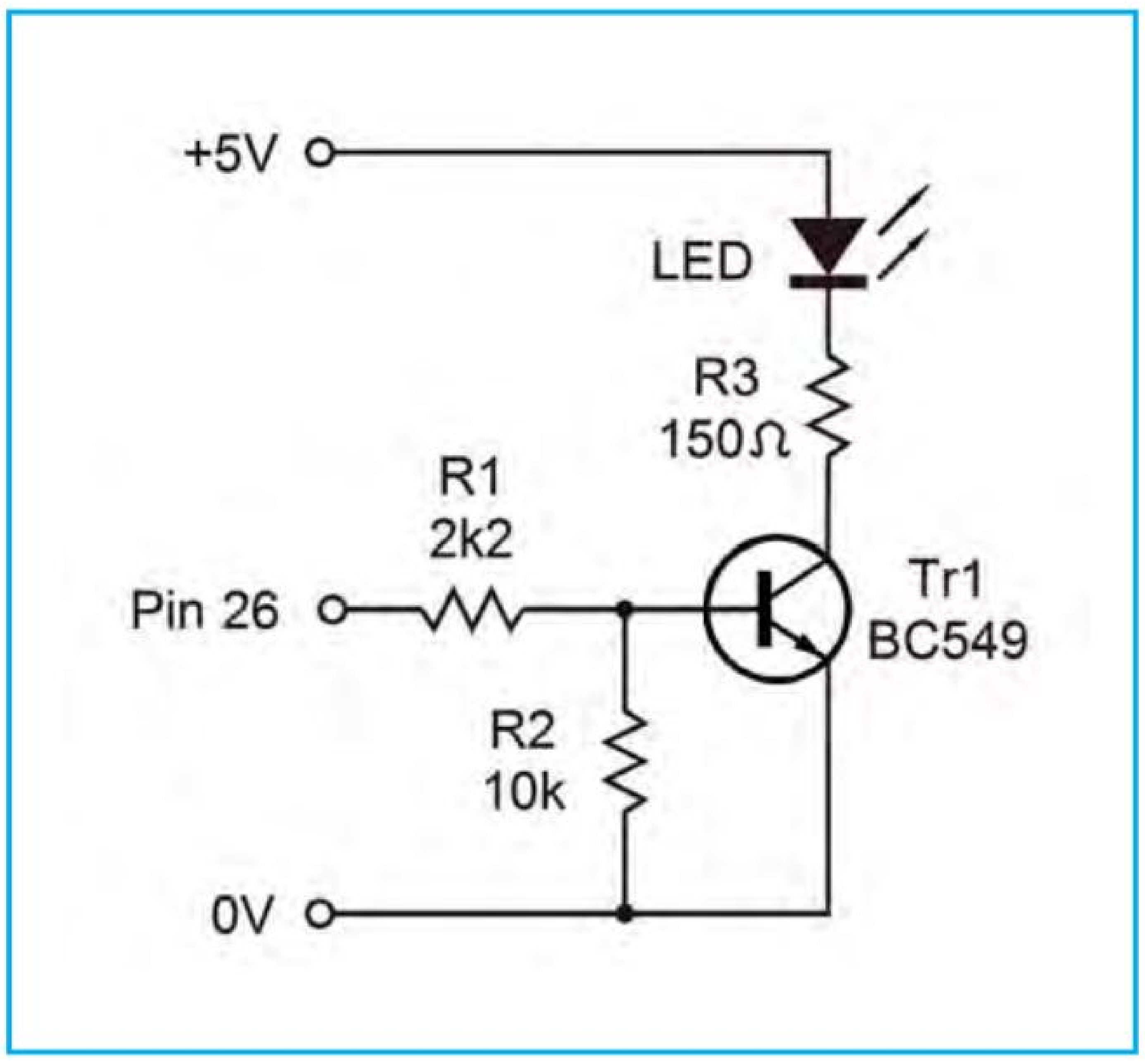


Fig.3. A simple LED driver circuit based on a common-emitter switch (Tr1). The 'on' current of the LED is set at about 20mA with the specified value for R3

to drive an LED (D1) via current-limiting resistor R3. The 'on' current is about 20mA. Any NPN transistor having a reasonably high current gain will suffice for Tr1, and any LED that operates in the visible light part of the spectrum will do for D1. Connection details for the 40-pin version of the GPIO port are provided in Fig.5. The 26-pin version is essentially the same, but it has pins 27 to 40 omitted.

Variable control

For practical applications it is necessary to have an easy way of varying the output signal. The program of Listing 2, in conjunction with the extra hardware of Fig.4, provides control of the output signal by way of three pushbutton switches. These switches are normally open types that do not latch when operated. The initial part of the program sets pin 26 of the GPIO port as the output for the PWM signal, but it also sets pins 18, 22 and 24 as inputs that have inernal pull-down resistors. Each of these inputs is therefore at the low state under normal conditions, but an input can be taken high by operating the appropriate switch. The duty cycle is set at zero initially, or continuously low in other words. The duty cycle of the PWM signal is controlled by the value in the variable called `dutcyc`, and this is set at a starting value of zero. Variable x is set at a value of 1 initially, and this is used to control, a `while`... `loop` that operates until x is set to a different value. The word `Starting` is printed on the screen to confirm that the program is going into the main routine.

The loop checks the states of the three input lines in turn using `if`... statements. If pin 18 goes high, x is made equal to 2. The loop then terminates and the PWM signal is switched off. Pin 22 going high results in `dutcyc` being increased by one on each loop of the program, gradually ramping it up to a value of 100. Taking pin 24 high has the opposite effect, with `dutcyc` being reduced by one on each loop. Two additional `if`... statements provide error checking that prevent the value in `dutcyc` going out of range. A potential problem with the program is that it would loop so fast that even momentary operation of S2 or S3 would almost instantly set the output at maximum or minimum. A `time.sleep` instruction is therefore used to provide a 0.1s delay in each loop so that the speed at which the output changes is more controllable. More on PWM next time, including more powerful driver circuits.

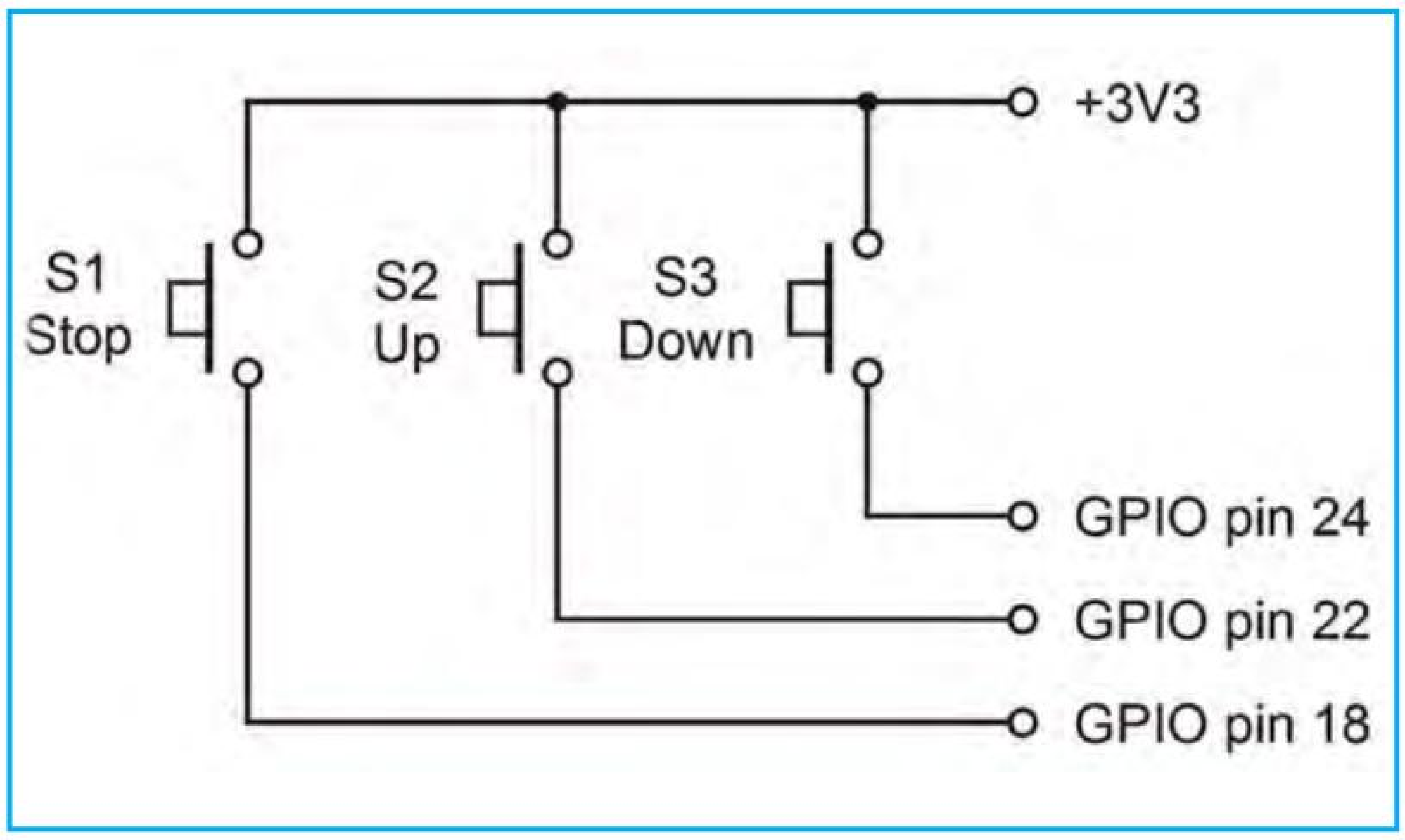
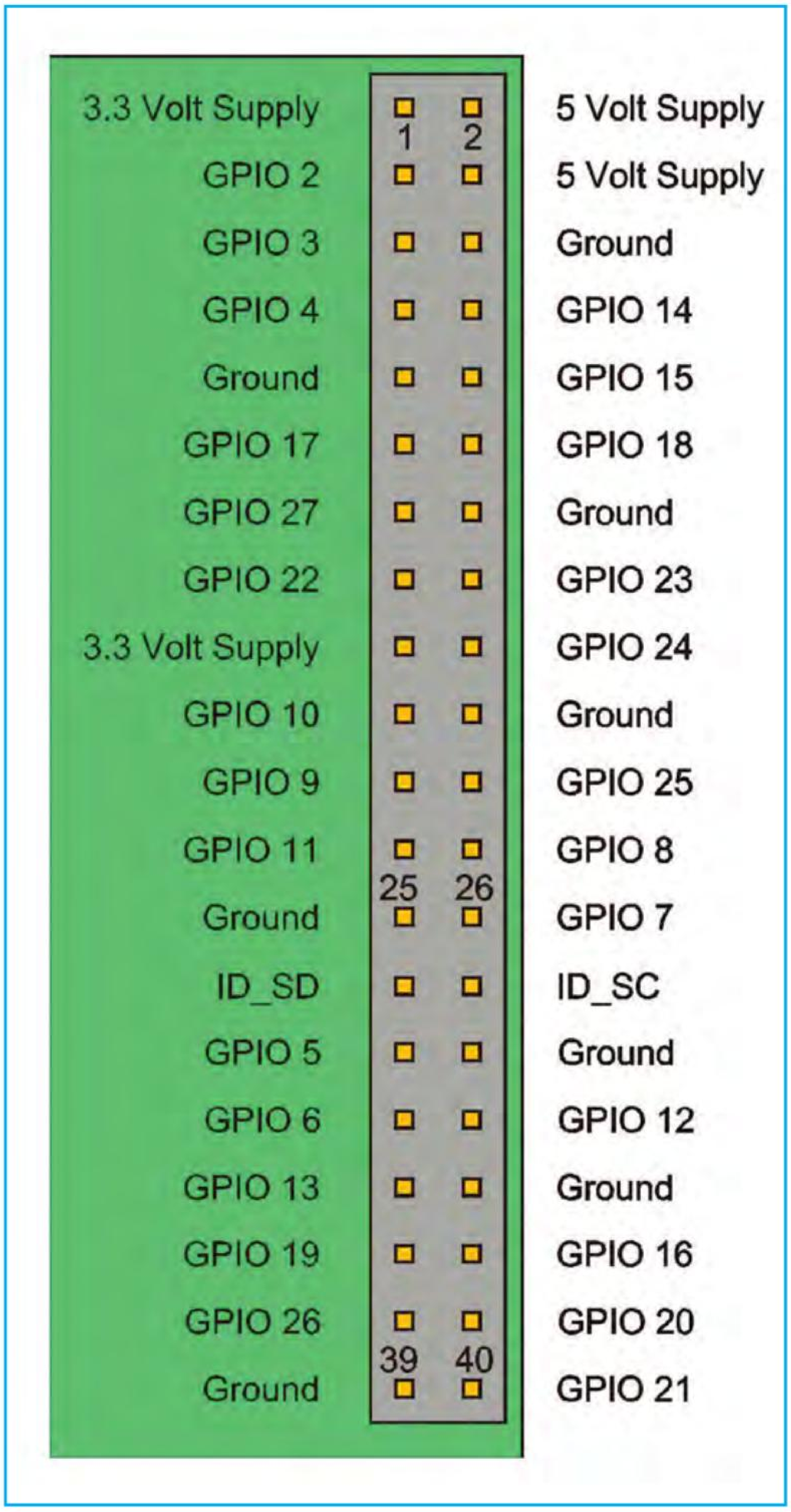


Fig.4. Three pushbutton switches plus suitable software can provide control of the PWM signal. S1 to S3 are simple, normally open, non-locking pushbutton switches



3.3 Volt Supply	1	2	5 Volt Supply
GPIO 2			5 Volt Supply
GPIO 3			Ground
GPIO 4			GPIO 14
Ground			GPIO 15
GPIO 17			GPIO 18
GPIO 27			Ground
GPIO 22			GPIO 23
3.3 Volt Supply			GPIO 24
GPIO 10			Ground
GPIO 9			GPIO 25
GPIO 11			GPIO 8
Ground	25	26	GPIO 7
ID_SD			ID_SC
GPIO 5			Ground
GPIO 6			GPIO 12
GPIO 13			Ground
GPIO 19			GPIO 16
GPIO 26			GPIO 20
Ground	39	40	GPIO 21

Fig.5. Connection details for the 40 pin version of the GIO port. The 26 pin version is the same, but with pins 27 to 40 omitted

CIRCUIT SURGERY

REGULAR CLINIC

BY IAN BELL

Current mirrors and transistor matching

CURRENT source circuits seem to be a popular topic this year; a few months ago we looked at op-amp-based current sources and also at the output characteristics of transistor current sources. User ***james*** posted the following question on the *EPE Chat Zone* forum.

'I recently breadboarded the discrete current mirror circuit in Fig.1. My understanding of a current mirror is that a current, I_1 is programmed by R1, which because of the identical V_{be} values of the two transistors, should lead to an identical current, I_2 in the second transistor. My results are as follow:

R2 = 100Ω, V_{ce} = 4.10V, I_2 = 59.0mA
R2 = 470Ω, V_{ce} = 3.37V, I_2 = 14.1mA
R2 = 680Ω, V_{ce} = 2.22V, I_2 = 11.5mA
R2 = 1000Ω, V_{ce} = 0.80V, I_2 = 9.2mA

'I_2 varies significantly with V_{ce} and is only similar to I_2 when the two V_{ce} values are similar. I selected two transistors that have similar values of h_{FE} measured on a meter with an h_{FE} measuring facility. I appreciate that h_{FE} varies slightly with V_{ce}, but this effect is too small to cause the wide variation in I_2 that I am seeing. It appears, at first sight, that using discrete transistors like this is not a practical way to build a current mirror – or am I doing something wrong? Why am I seeing such a wide variation in I_2 as the load (R2) varies? My interest in this area was sparked by Robert Penfold's use of a discrete current mirror in August's *Interface* article.'

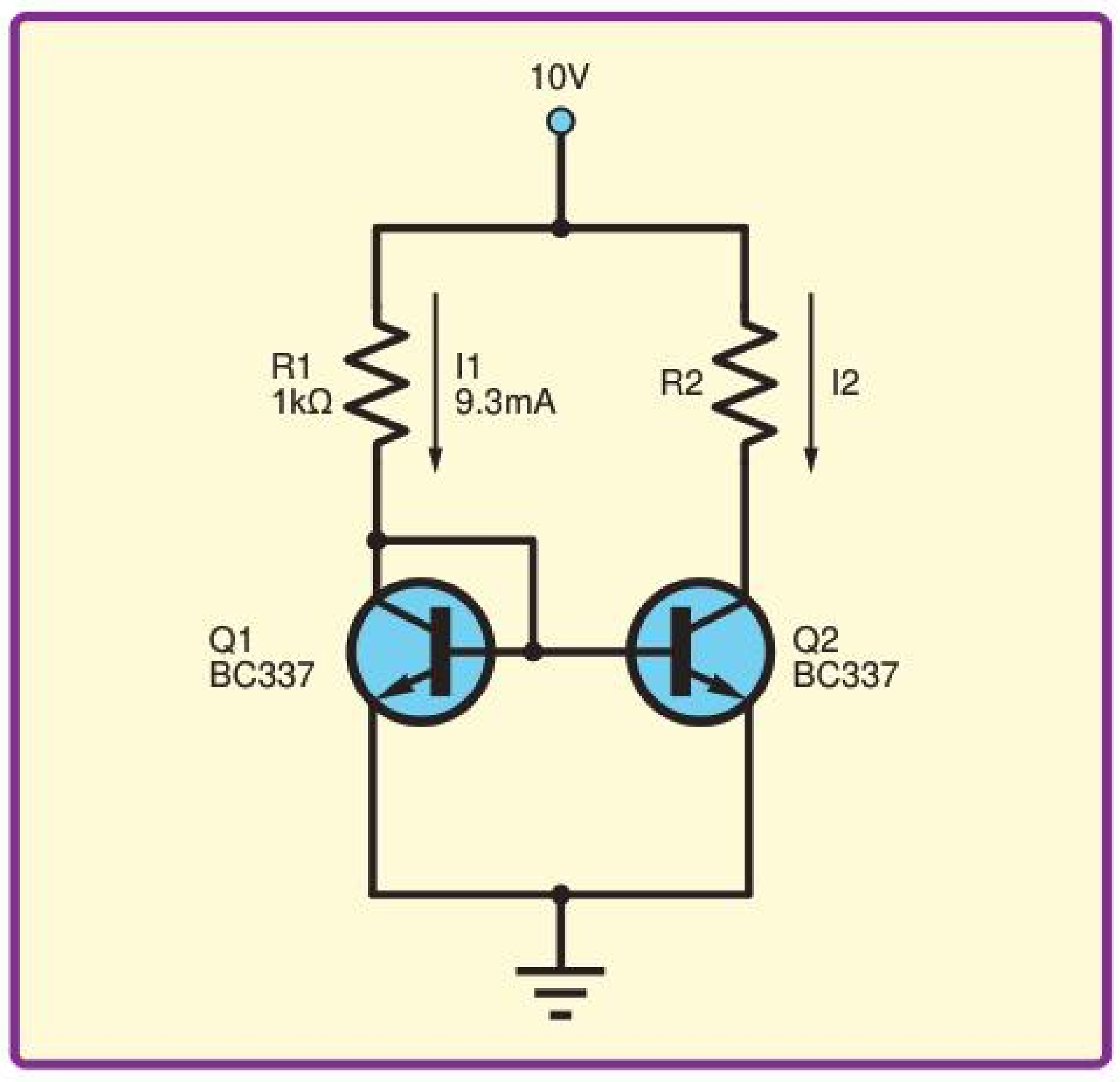


Fig.1. ***james****' current mirror circuit*

james's questions raises some issues not discussed in previous articles, particularly in relation to the accuracy of the current mirroring behaviour of the circuit. The current output (I_2 in Fig.1) is a copy of the reference current (I_1 in Fig.1). If the circuit was a perfect current mirror then these two currents would always be exactly equal, but of course in a real circuit this may not be the case. The circuit's output (from the collector of Q2 in Fig.1) acts as a current source. If I_1 is constant, which it should be if R1 and the supply do not change, then Q2 acts as a constant current source.

Before looking at ***james's*** question in more detail, we will briefly recap/introduce current sources and explain the current mirror operation of the circuit in Fig.1. For further background on current sources, please look at *Circuit Surgery* in the March 2015 issue of *EPE*.

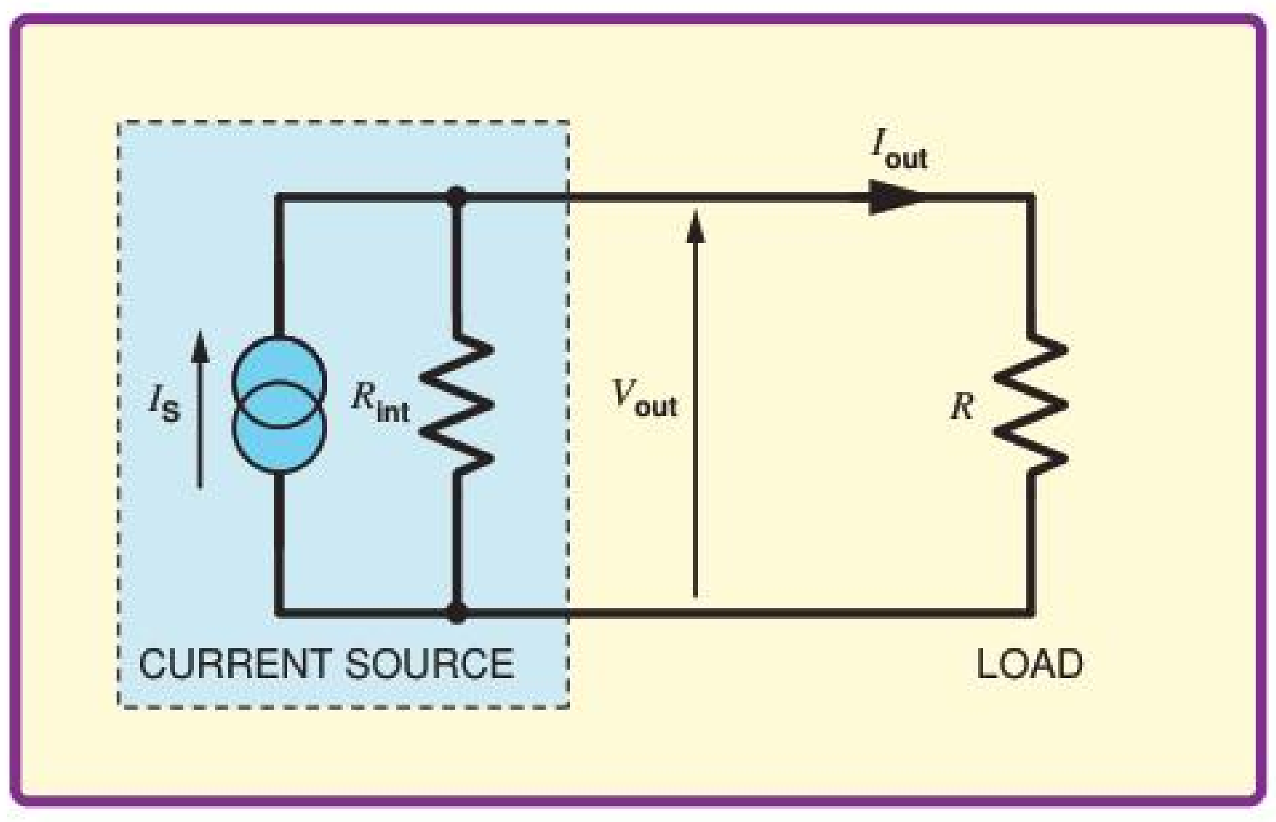


Fig.2. Internal resistance of a real current source

Current sources

An ideal current source outputs a particular current irrespective of the voltage across it. Real current sources have internal resistance (see Fig.2), which can be modelled as being in parallel with an ideal current source. The larger the internal resistance, the better the current source. Compare this with a voltage source, where the internal resistance is in series and the smaller the resistance, the more ideal the source.

If we connect a large resistance across an ideal current source, then by Ohm's law it must develop a large voltage across the resistor. For example, if we connect a 1MΩ resistor across a 1mA current source it would have to produce 1000V across the resistor to maintain the required current, impossible in the circuit of in Fig.1 due to the limited supply voltage.

The internal resistance, R_{int}, of an otherwise ideal current source, I_S, limits its output voltage, V_{out}, to $I_S R_{int}$ (Fig.2), at which point I_{out} is zero – so this is the open-circuit output voltage. The larger the value of R_{int}, the larger the maximum output voltage. For truly ideal current sources, the open-circuit output voltage is infinite and R_{int} is infinite – these do not exist in reality, but are a useful mathematical model. If we short-circuit a current source (R = 0 in Fig.2) the output current is equal to I_S and there is zero voltage across the source.

Real current sources also have a limited range of voltages over which they can operate (called 'compliance'), often determined by factors other than the internal resistance, for example when the transistors in the circuit switch off, change operating mode as their bias conditions change, or reach their operating limits.

Current mirrors

In the circuit in Fig.1 the collector and base of Q1 are shorted together; so we are left with the base-emitter PN junction – the transistor acts as a diode (Fig.3a). We can bias this diode to carry any reasonable forward current by connecting a resistor from the supply, as shown in Fig.3b. The forward voltage drop of the 'diode' will be the V_{BE} of the transistor with the collector current set by the resistor.

If we wire this diode-connected transistor to another transistor –

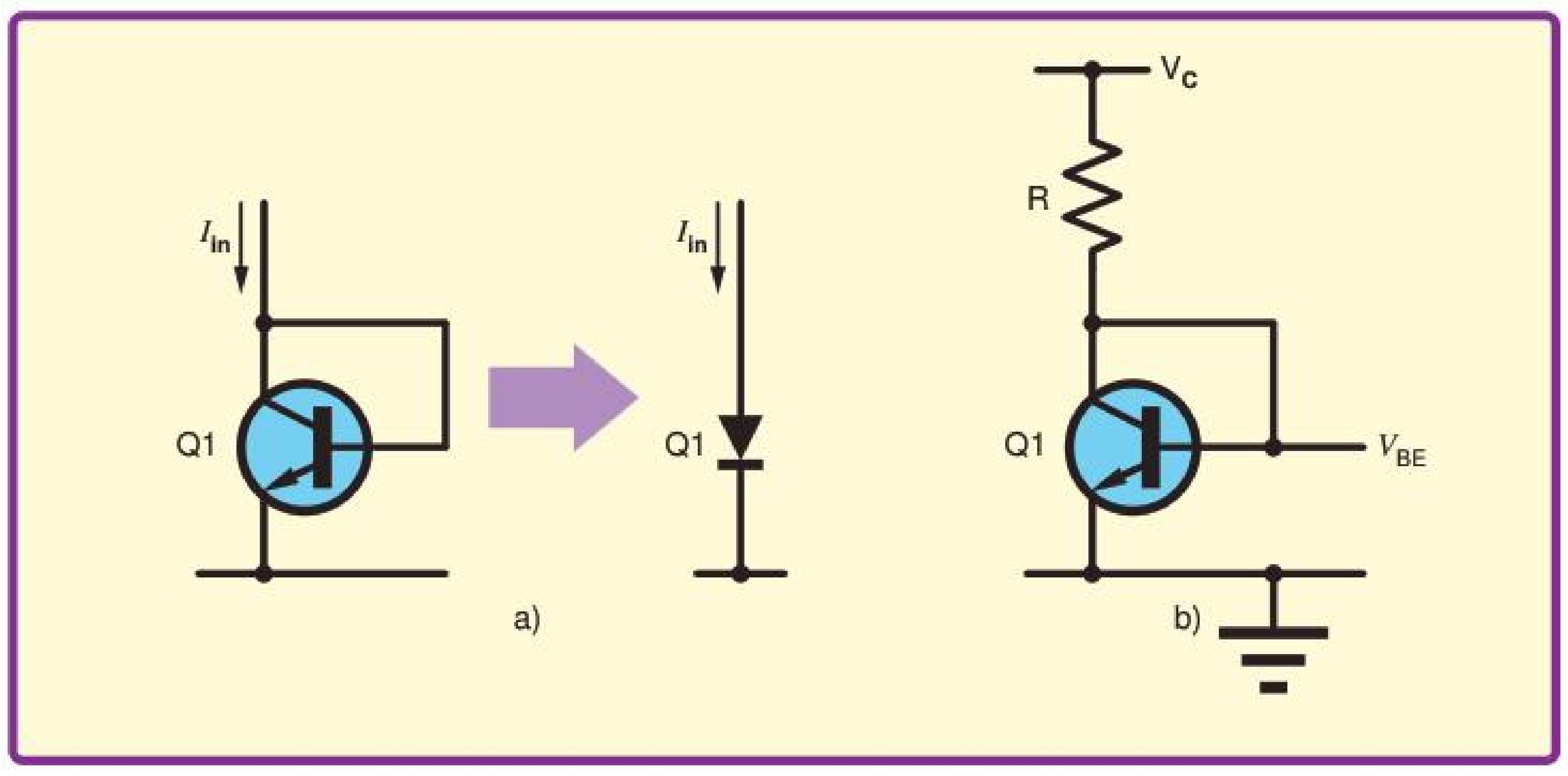


Fig.3. Diode-connected transistor

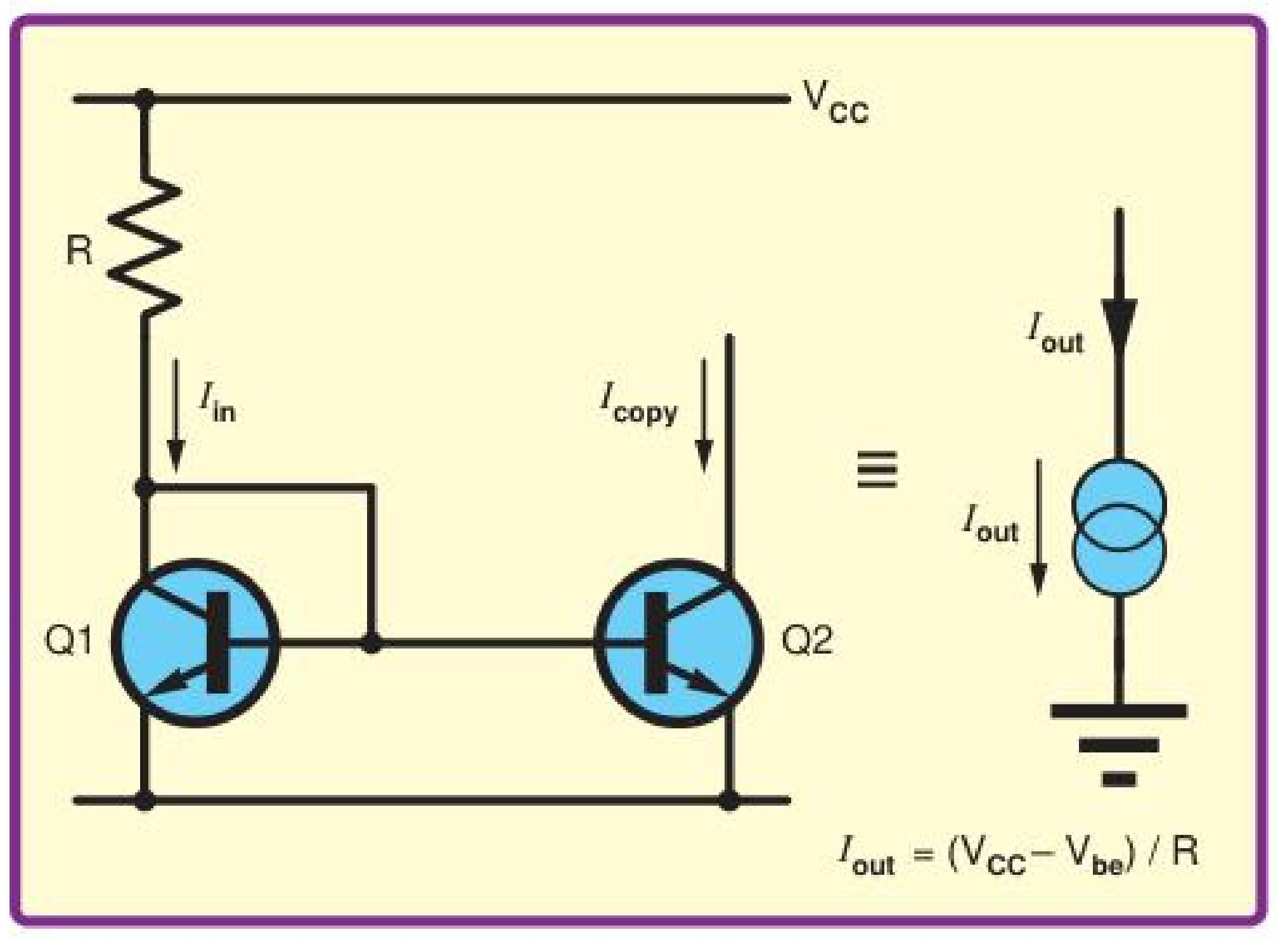


Fig.4. The current mirror

emitter-to-emitter and base-to-base – they will both have the same V_{BE} (see Fig.3). If the transistors have equal characteristics, then whatever current is set for the first transistor will also flow in the second ($I_{copy} = I_{in}$ in Fig.4). This is a basic current mirror circuit.

To get a constant current source (rather than a current mirror) we supply a fixed input current through the diode-connected transistor, as in Fig.1. To calculate the value of R we can assume that the base-emitter voltage of the transistor V_{BE} is fixed, and in the range 0.6-0.7V, the output current is then:

$$R = (V_{CC} - V_{BE}) / I_{out}$$

If we have more details of the transistor, such as the V_{BE} *vs* I_C curve then we can set this more accurately.

Current mirrors are very important in electronic circuit design; they are to be found inside almost every complex analogue IC (such as operational amplifiers) where they have widespread general use for biasing and as active loads.

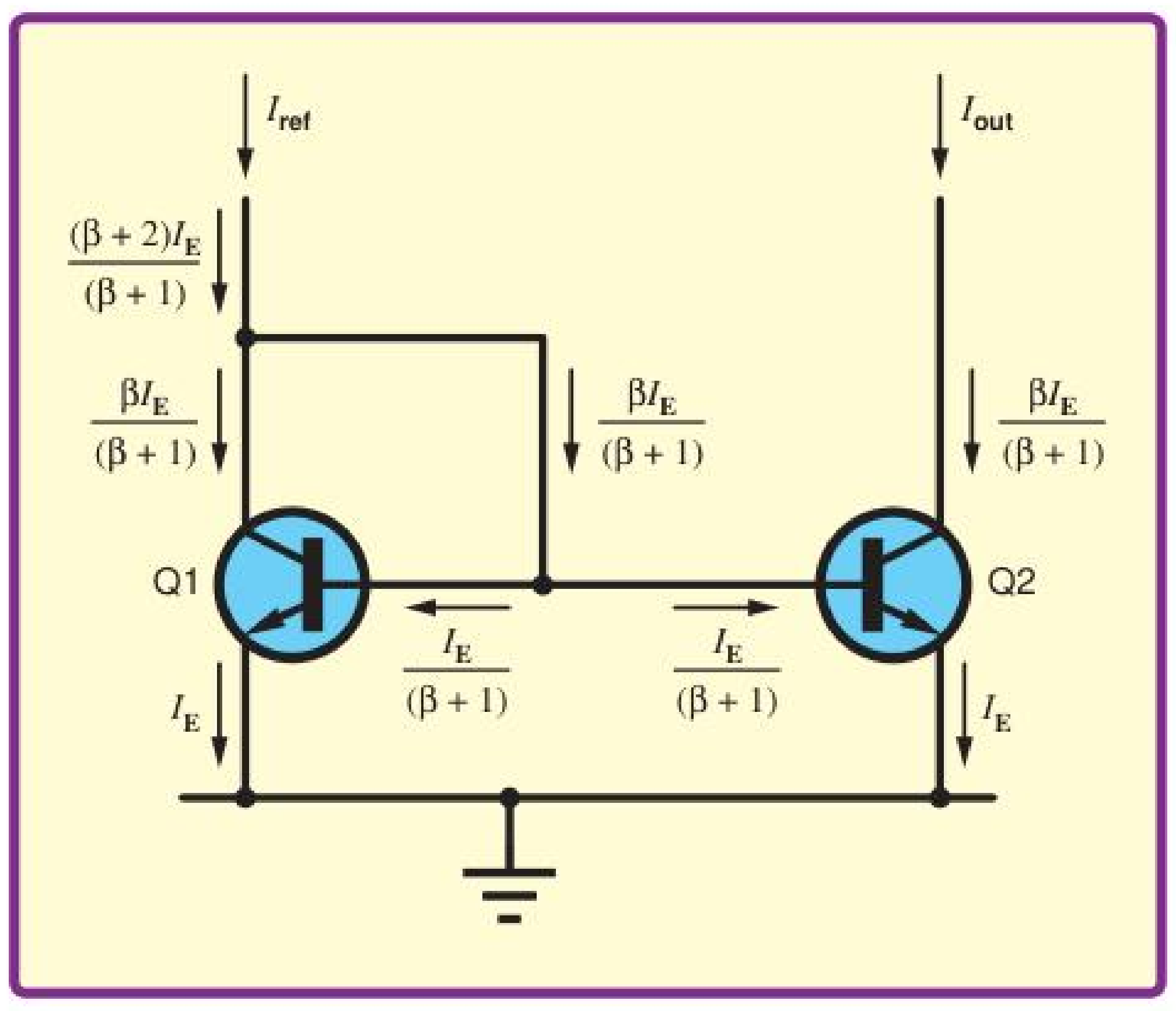

I_{ref}
I_{out}
$\frac{(\beta+2)I_E}{(\beta+1)}$
$\frac{\beta I_E}{(\beta+1)}$
$\frac{\beta I_E}{(\beta+1)}$
$\frac{\beta I_E}{(\beta+1)}$
Q1
Q2
$\frac{I_E}{(\beta+1)}$
$\frac{I_E}{(\beta+1)}$
I_E
I_E

Fig.5. Currents in the basic current mirror

Mirror accuracy

Fig.5 shows an analysis of currents in the basic current mirror. This starts with the assumption that the transistors are exactly equal in terms of both characteristics and environment. So we can assume they both have the same forward current gain β (= h_{FE}) and just use β in our calculations, rather than distinguishing, say β_1 and β_2.

Given that the two transistors have the same V_{BE} (from the way the circuit is wired), they must also have the same I_E (because the characteristics are the same). So currents in the circuit can be determined using $I_C = \beta I_B$ and $I_E = (1 + \beta)I_B$, together with application of Kirchoff's current law at the wire junctions. Thus, from Fig.5 we have:

$$I_{ref} = \left(\frac{\beta+2}{\beta+1}\right) I_E$$

and

$$I_{out} = \left(\frac{\beta}{\beta+1}\right) I_E$$

so

$$I_{out} = \left(\frac{\beta}{\beta+2}\right) I_{ref} \approx I_{ref}$$

That is I_{out} is approximately equal to I_{ref} if β is much greater than two. For a 'rule of thumb', value of β of 100, the output is about 98% of the reference current.

We don't know what gain ***james's*** transistors had, but the BC337-25 model in the LTSpice circuit simulator (which we will use shortly) has β = 292.4. Using this in the above equation with I_{ref} = 9.3mA from Fig.1 we get I_{out} = 9.2mA, which is consistent with ***james's*** result for the 1kΩ load. Of course, this does not explain the other measured values.

Simulation set up

We can simulate the circuit in Fig.1 in LTSpice using the circuit in Fig.6. Before showing the results we will discuss some details of how this simulation is set up. The simulation we are performing is an operating point analysis. This simply calculates the DC currents and voltages in the circuit – capacitors are treated as open circuits and inductors as short circuits; we do not get any waveforms or frequency response curves. The output from SPICE is just a text listing of the DC currents and voltages.

In Fig.6 we have extended the basic operating point calculation by adding a parametric sweep, that is, we instruct LTSpice to run the same simulation many times with different values of one or more component values. In this case we are using values of R2 from 2Ω to 1000Ω in steps of 5Ω, as defined by the Spice directive, `.step param R 2 1000 5`

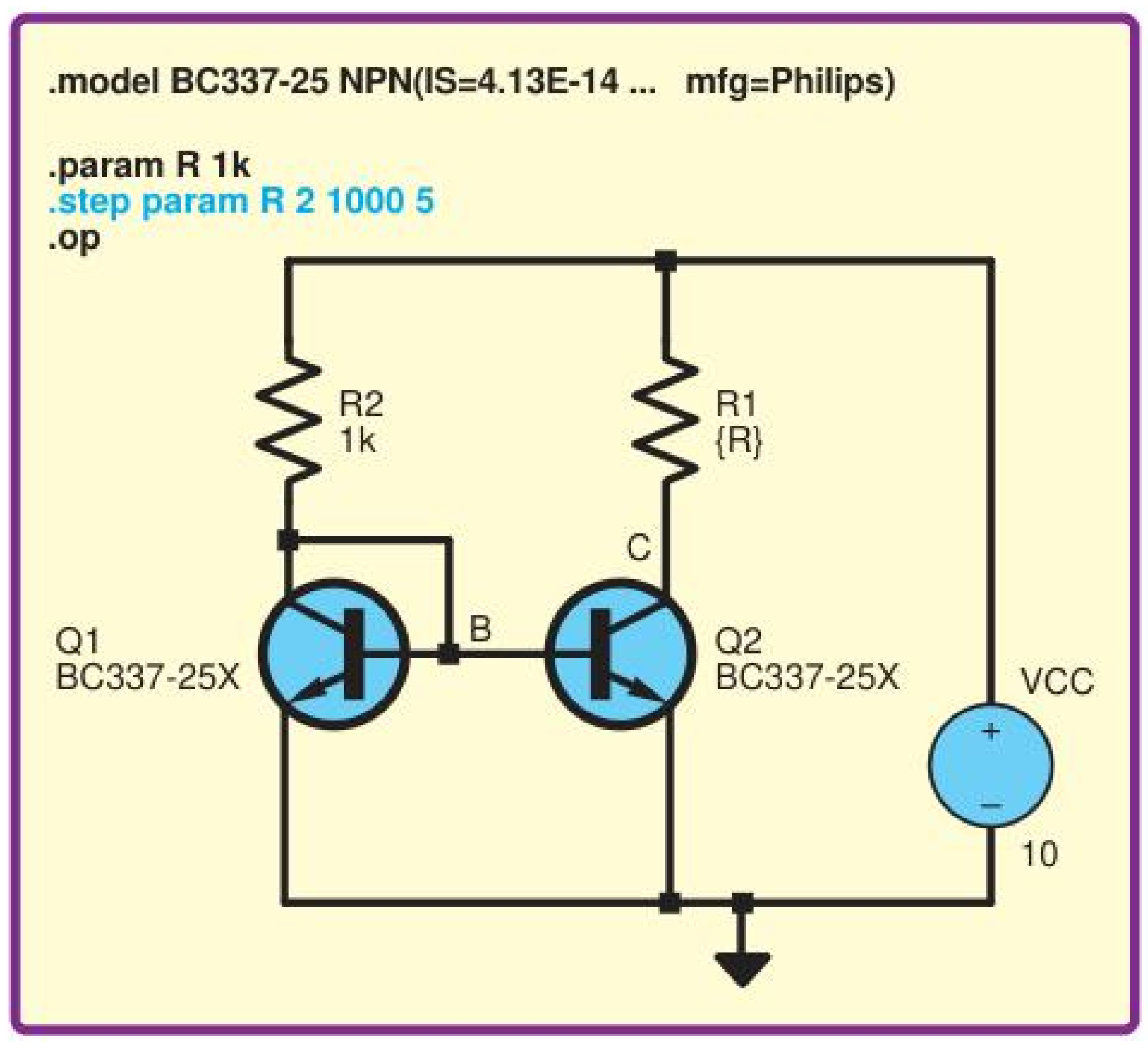


Fig.6. LTSpice schematic for simulating Fig.1

The reistance parameter R is declared by the SPICE directive, `.param R 1k`, which sets its default value to 1kΩ. The value of R2 is set to `{R}` which is SPICE syntax for referring to a parameter value. Using a parameter sweep allows us to plot a graph of other circuit values (eg, current) against the parameter value.

We can easily switch between running a simple operating point simulation and the full parameter sweep by right clicking on the `.sweep` directive and selecting 'Comment' rather than 'SPICE directive'. The `.sweep` text is shown blue when it is just a comment and R2 takes the default value of parameter R.

Models

Another point to note about the schematic in Fig.6 is that we have include a SPICE `.model` statement directly on the schematic. This model provides characteristics for the transistor (in this case, the BC337 used by ***james***). LTSpice provides this model as part of its download. When you right click a bipolar transistor you can select BC337-25 or BC337-40 (versions with different gains) from the list of real transisor models provided with LTSpice. The models are defined in a text file, **standard.bjt**, in the lib folder tree, within the LTSpice installation. Typically, the location will be something like: **C:\Program Files (x86)\LTC\LTspiceIV\lib\cmp**.

Here, the `.model` directive was simply copied from the BC337-25 `.model` directive in **standard.bjt**, pasted into a SPICE directive, the model name was changed to BC337-25X and the directive placed on the schematic. The transistor type value was set to BC337-25X to match the model name. This approach allows us to change the model parameters in our simulations and quickly have access to the full model. The full model text is not shown on the schematic, it is:

```
.model BC337-25
NPN(IS=4.13E-14 NF=0.9822
ISE=3.534E-15 NE=1.35
BF=292.4 IKF=0.9 VAF=145.7
NR=0.982 ISC=1.957E-13 NC=1.3
BR=23.68 IKR=0.1 VAR=20 RB=60
IRB=2.00E-04 RBM=8 RE=0.1129
RC=0.25 XTB=0 EG=1.11 XTI=3
CJE=3.799E-11 VJE=0.6752
MJE=0.3488 TF=5.4E-10 XTF=4
VTF=4.448 ITF=0.665 PTF=90
CJC=1.355E-11 VJC=0.3523
MJC=0.3831 XCJC=0.455
TR=3.00E-08 CJS=0 VJS=0.75
MJS=0.333 FC=0.643 Vceo=45
Icrating=500m mfg=Philips)
```

Simulation

The results from the LTSpice sweep simulation are shown in Fig.7. Here we see that the currents are generally not equal and the output current varies with the value of the load, R2.

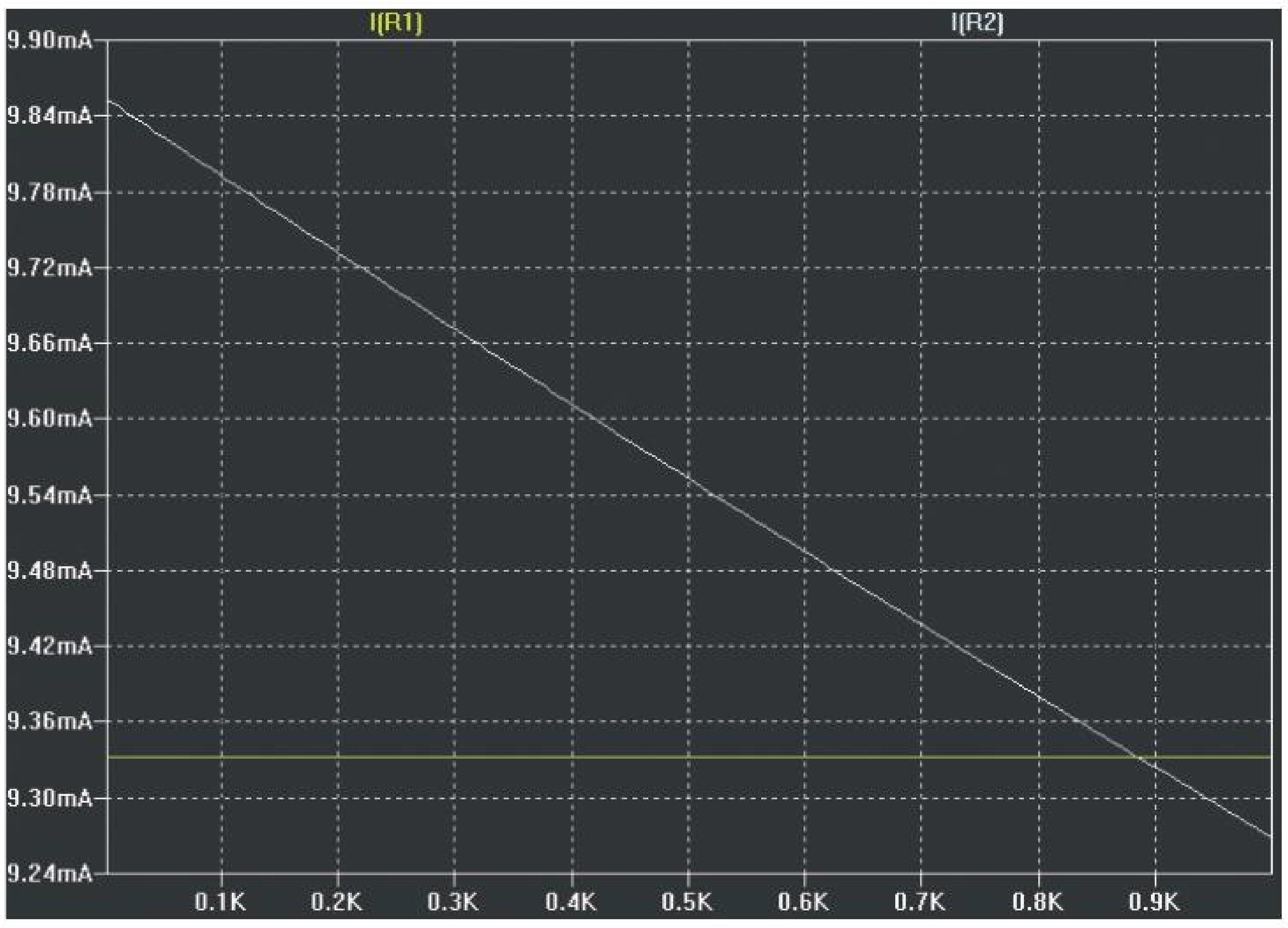


Fig.7. Currents in R1 and R2 in the circuit in Fig.6 against the value of R2 in ohms

Ideally, the output current should be independent of R2 – if the current mirror output behaved as a true constant current source. The non-ideal behaviour is due to the Early effect in transistor Q2, the effect of which is for the transistor to behave as a current source with non-infinite output resistance (see discussion relating to Fig.2 earlier). We discussed this aspect of current sources in more detail in *Circuit Surgery* in *EPE* May 2015.

At the moment, we are interested in other aspects of the transistor and can idealise the transistor model with respect to output resistance by setting the Early voltage to zero. To do this we can edit the `.model` directive and change `VAF=145.7` to `VAF=0`. Resimulating with the Early voltage set to zero gives the results shown in Fig.8.

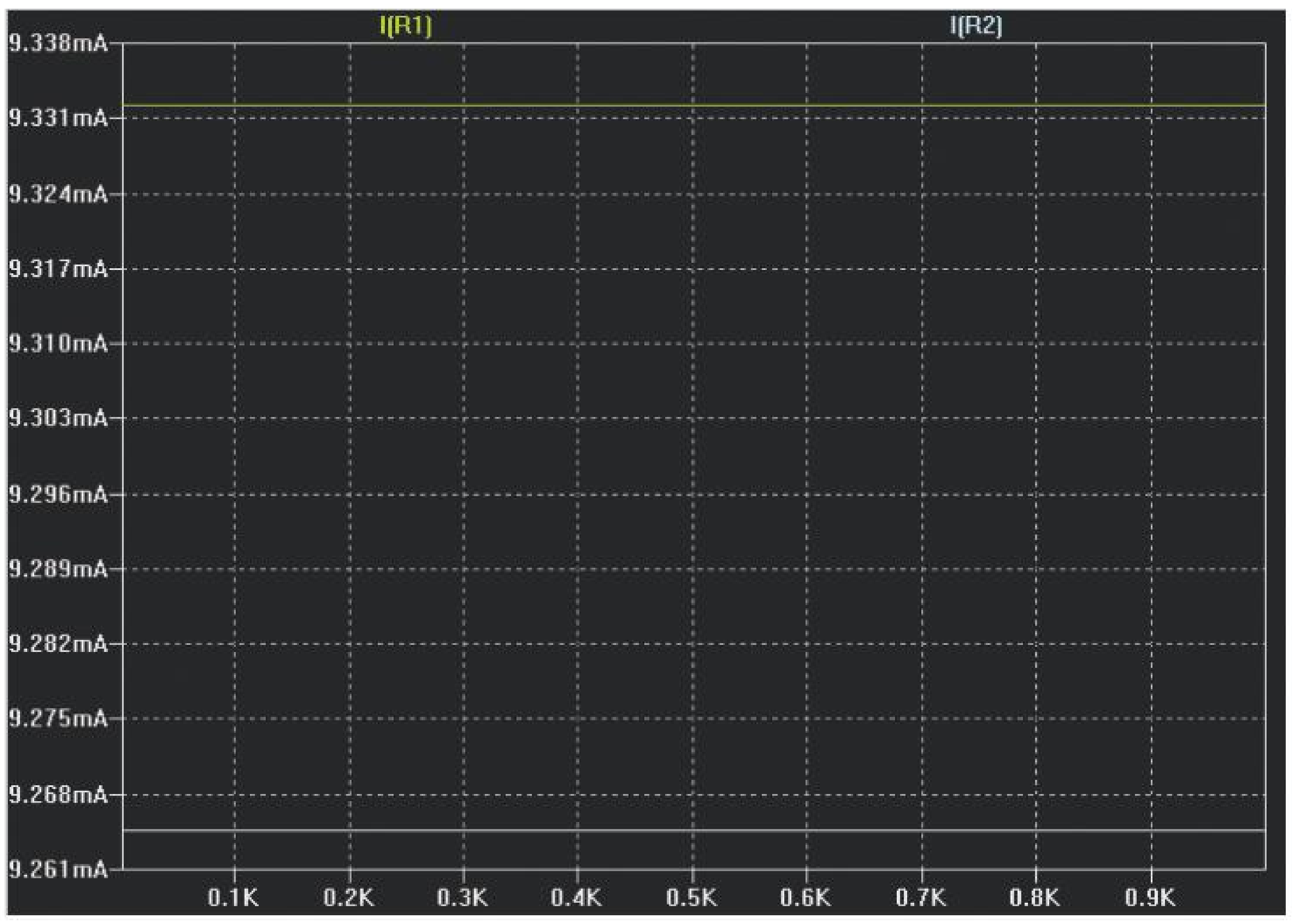


Fig.8. Currents in R1 and R2 in the circuit in Fig.6 against the value of R2 in ohms with the transistor Early voltage set to zero

The current values now do not change and simulating the model without the sweep gives the following results, which is more convenient for reading the numerical values.

```
     --- Operating Point ---

V(c):          0.73537        voltage
V(b):          0.667799       voltage
V(supply):     10             voltage
Ic(Q1):        0.00926463     device_current
Ib(Q1):        3.37854e-005   device_current
Ie(Q1):        -0.00929842    device_current
Ic(Q2):        0.00926463     device_current
Ib(Q2):        3.37854e-005   device_current
Ie(Q2):        -0.00929842    device_current
I(R1):         0.0093322      device_current
I(R2):         0.00926463     device_current
I(Vcc):        -0.0185968     device_current
```

This matches with our calculations in Fig.5 – the two transistors have exactly the same base, emitter and collector currents, but the resistor currents are slightly different. Putting $I_{\text{ref}} = 0.0093322$ and $\beta = 292.4$ into the current mirror equation above gives 0.00926880, very similar to the R2 current listed above.

Mismatch

We now have a simulation of the current mirror working well, but no explanation for the high currents observed by ***james***. One aspect of reality, which is important here, but which has not been included in the simulation is the fact that the two transistors are not identical. In the simulation, the transistors both have the same model (BC337-25X) and hence exactly the same characteristics. We can fix this by giving them different models. We can cut and paste the existing `.model` directive on the schematic to create a copy, and then edit the two to have different names (eg, BC337-25X1 and BC337-25X2), then we edit the transistor type names so each has a different one of the two new model names. After this we can change the parameters for the individual transistors and resimulate.

If we set the gain for Q2 to 100 (BF=100 in the model) we get the following results:

```
     --- Operating Point ---

V(c):          1.47173        voltage
V(b):          0.66763        voltage
V(supply):     10             voltage
Ic(Q1):        0.00920909     device_current
Ib(Q1):        3.35816e-005   device_current
Ie(Q1):        -0.00924268    device_current
Ic(Q2):        0.00852827     device_current
Ib(Q2):        8.9692e-005    device_current
Ie(Q2):        -0.00861796    device_current
I(R1):         0.00933237     device_current
I(R2):         0.00852827     device_current
I(Vcc):        -0.0178606     device_current
```

Here we see that the transistors' base, collector and emitter currents are no longer equal and the R2 current is not so close to the R1 current. Observant readers may have realised that we could contrive to make the resistor currents equal by selecting exactly the right gain for Q2, but this is not really a constructive exercise because we do not have such precise control over the gain of our real transistors.

Getting hotter

Another observation from experimenting with varying Q2's gain is that we cannot achieve the levels of current observed by ***james*** with realistic values of gain, so it is not gain mismatch which is the cause of the high currents he observed. We stated earlier in our analysis of the current mirror that we assumed that the transistors are exactly equal in terms of both characteristics and environment. Usually, the most important environment parameter is temperature.

SPICE sets a default temperature of 27°C for all components unless otherwise instructed. We can set the temperature for all components using the SPICE directive `.OPTION temp=value` but in this case we need to set the temperature of an individual transistor. To change the temperature of a transistor we right click the type/model name text and add `temp=value` after the model name, for example, for Q2, with model BC337-25X2 at 65°C the transistor type label becomes `BC337-25X2 temp=65`

If we set the transistor gains back to their default value, but keep the Early voltage at zero and resimulate using a sweep of R, as before, but with Q2 at 65°C we get the results shown in Fig.9. There are a couple of key things we need to discuss regarding these results.

Results

First, we can see that the current values are not exactly the same as ***james's*** values, but they are consistent with

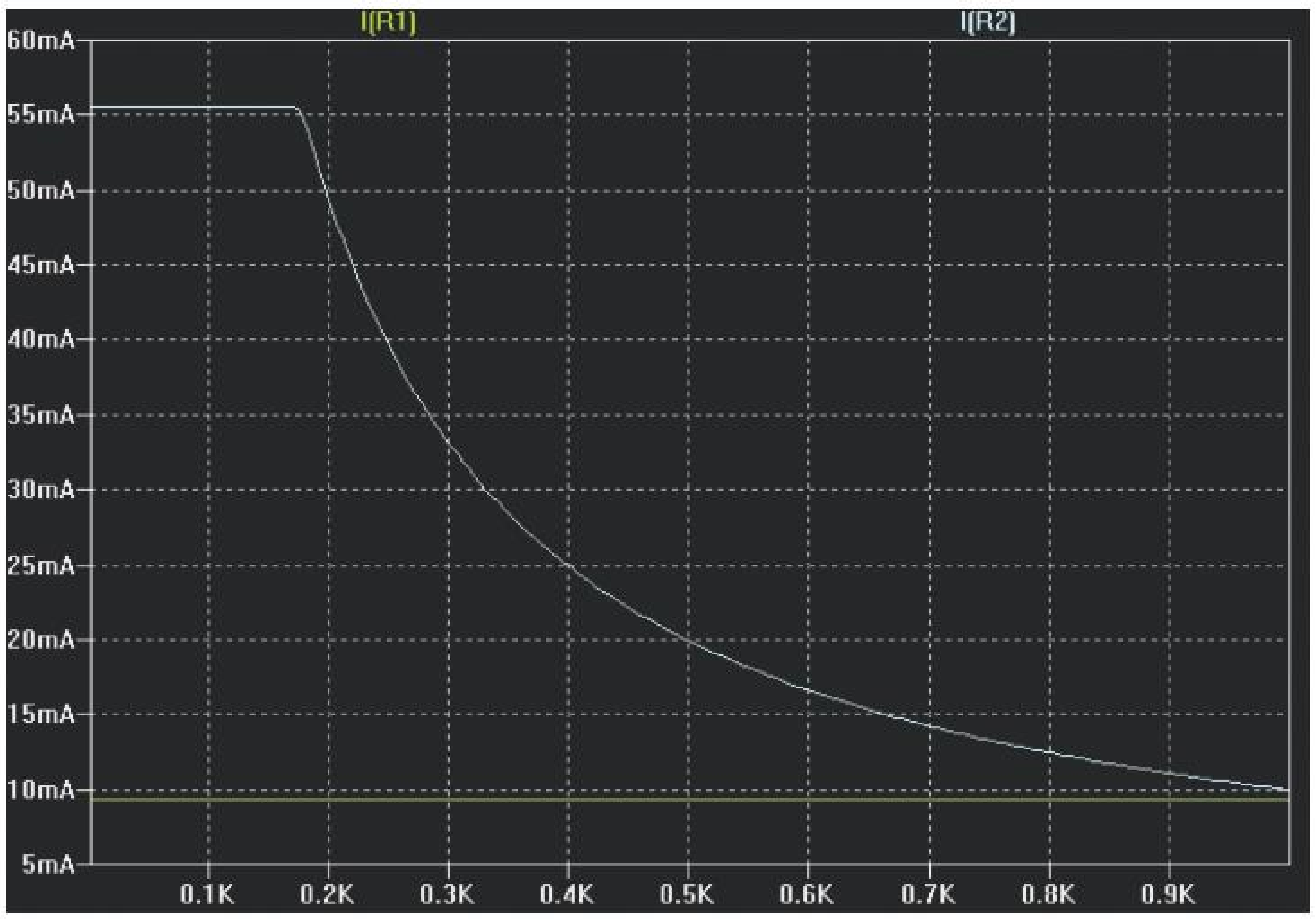


Fig.9. Currents in R1 and R2 in the circuit in Fig.6 against the value of R2 in ohms with the transistor Early voltage set to zero. Temperature of Q2 is 65°C

them (we don't know the precise value of gain and other parameters of his transistors or their temperatures, so we cannot expect to be exact). This indicates that a mismatch in transistor temperature is probably the cause of the poor current mirroring observed by ***james***. The idea that temperature mismatch was the probable cause was raised by another regular *Chat Zone* contributor, ***bowden_p***, in the discussion thread following ***james's*** post. If the reference current is set lower (***james*** reported on this later in the thread) the self-heating will be significantly lower, and the transistors are more likely to remain at more or less the same temperature.

Second, we can see in Fig.9 that the constant current behaviour of the circuit breaks down for resistor values above about 170Ω. This is due to Q2's collector-emitter voltage dropping below its saturation voltage (V_{CEsat}). The transistor is then not operating in forward-active mode and does not have constant current output as the collector-emitter voltage varies. The larger the current and the higher the value of R2, the greater the voltage drop across R2, and hence the lower the collector-emitter voltage. This issue was discussed in depth in the May 2015 *Circuit Surgery*, so we will not go into further details here.

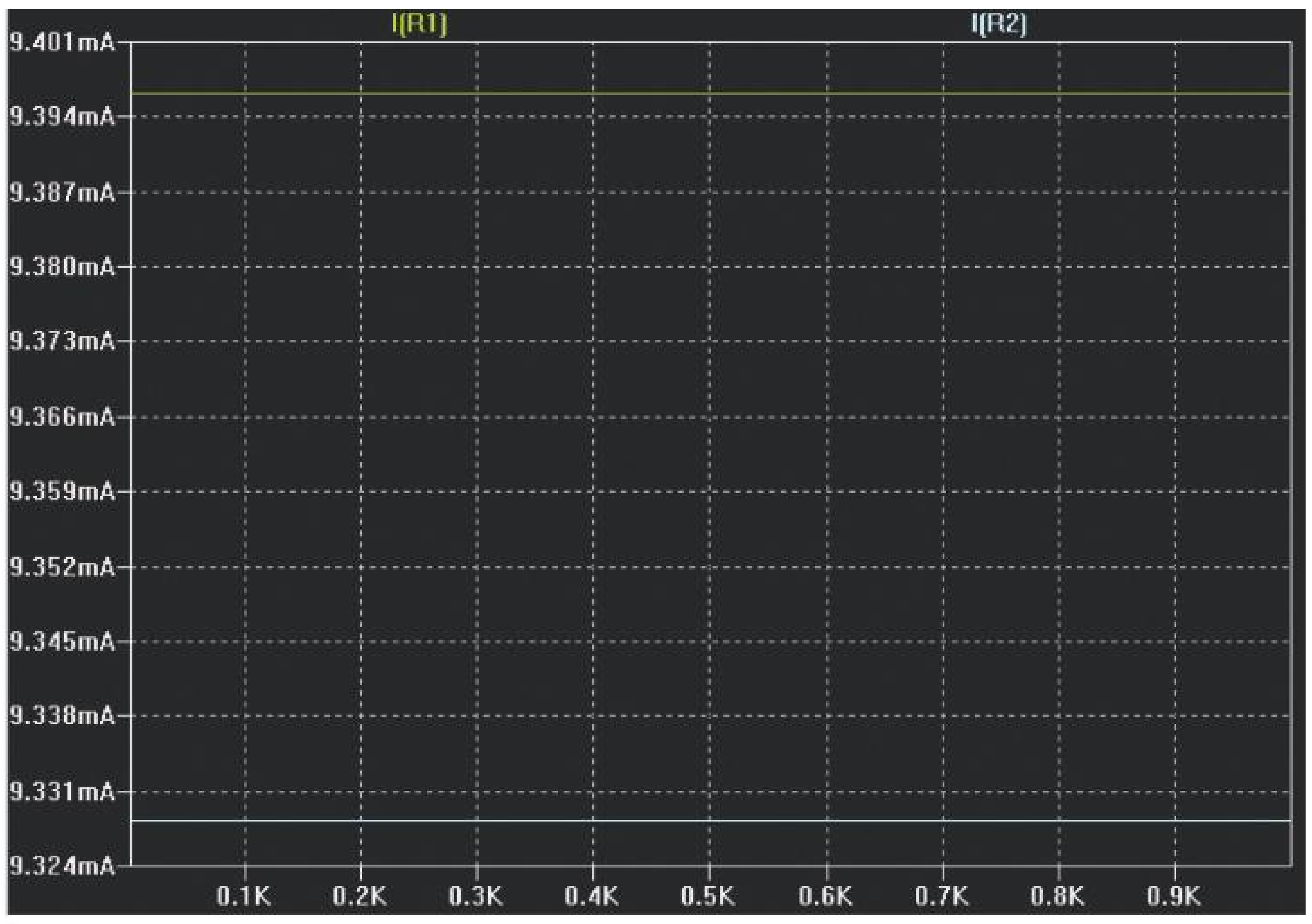


Fig.10. Currents in R1 and R2 in the circuit in Fig.6 against the value of R2 in ohms with the transistor Early voltage set to zero. Temperature of both transistors is 65°C

One point worth checking is that the circuit still works correctly with *both* transistors at an elevated temperature. If we set them both to 65°C we get the results shown in Fig.10. Here we see that the mirroring, and constant current output, is working as before, with only a slight increase in current in both transistors. If the transistors are both at the same temperature the mirror works correctly. In practical terms, with discrete transistors, we could clamp them both close together on the same heatsink to help achieve this.

An issue raised in the *Chat Zone* discussion was how does the mirror get into the state where one transistor is hotter then than the other, given that they start at the same temeprature at power up. The answer is that they both self-heat as they dissipate power, which will tend to cause the collector currents to increase. This current is ultimately limited by the collector load resistors – it is not possible for the collector current to be larger than the supply voltage divided by the load resistance. With R1 = 1kΩ and R2 = 100Ω these limits are 10mA and 100mA respectively. If the transistors are not held in thermal equilibrium, then the mirroring breaks down, and they are free to self-heat and increase their currents independly within these limits. If are they forced to have the same temperature then this cannot happen and the current in Q2 will be set by the current in Q1.

Matched

Matched transistors are often used in IC designs in current mirrors and other circuits. IC designers use layout techniques to ensure the transistors are closely matched, both in terms of charateristics and temperature. Two transistors close togeter on the same chip are likely to be at similar temperatures, but gradients in temperature across the chip are still possible. One approach to countering this, which is not applicable to discrete transistors, is to split each transisor into two parts and and place them on the silicon so that they have a common 'centre of gravity'; this is known as common centroid layout. Using this approach, any linear gradients (eg, in temperature and dopant levels) will have, on average, equal effect on both transistors, ensuring a close match.

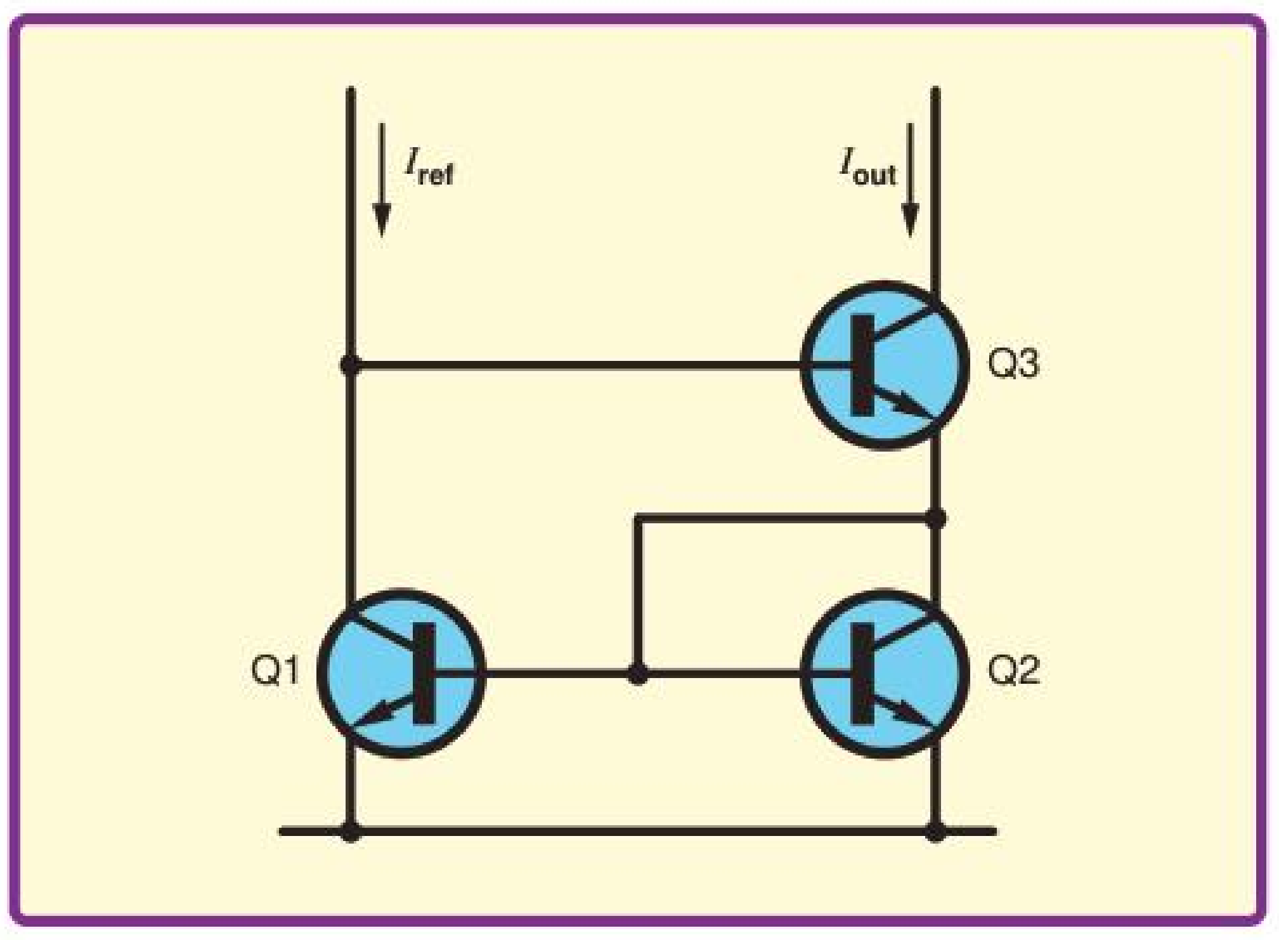


Fig.11. Wilson current mirror

Even if the transistors are well matched, the circuit in Fig.1 does not produce a very accurate copy of the reference current. However, there are better circuits available, for example the Wilson current mirror (see Fig.11), which uses three transistors and has better matching and better output charateristics (higher current-source internal resistance).

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PIC n' Mix

Mike O'Keeffe

Our periodic column for PIC programming enlightenment

Passing of the torch and concluding the LPLC Oscilloscope

THIS MONTH we say farewell and good luck to Mike Hibbett, who has written the *PIC n' Mix* column for over a decade. Mike has kindly passed the torch to me, and I am truly delighted and honoured to be taking over his position. I'm sure it's not the last we'll hear from Mike – I'm hoping to persuade him to do a guest article or two for us in the future. As a close friend, I wish him every success in his new ventures.

So, who am I? I'm an electronic engineer and I have loved building gadgets since the age of nine – building my very first radio with a lollipop stick attached to a ferrite rod, which moved up and down the centre of a poorly wound coil made from a Weetabix box. I have built, fixed and wrecked many gadgets in the pursuit of knowledge and fun. One of my latest designs is a Teslatronix Tesla Coil, which illuminates LEDs and energy saver bulbs up to a foot away using a 9V PPC battery.

To start off, I will be completing Mike Hibbett's PIC-based *LPLC Oscilloscope*, using the LPLC board developed from his very successful Kickstarter campaign.

Let's get a quick reminder of what the project involved. First, we looked at developing our own microcontroller circuit with a custom-made PCB back in January 2014. After a reader request, the idea of starting a Kickstarter project was conceived, resulting in a very successful campaign. A total of 430 LPLC boards were sold and shipped to those who supported the campaign (including me).

I used my LPLC board for a test system in manufacturing. It had a number of ADC inputs to monitor signals and voltages under varying conditions. When everything was good, a number of green LEDs were illuminated, when anything was bad, the corresponding red LED would light – nice, simple and effective.

Back to our PIC Oscilloscope

What's next? We've hooked up the screen, power, ground, our SPI interface, the reset and other signals. We've programmed the PIC and we see the grid on the screen, with a green horizontal line representing our signal.

We want to keep our *LPLC Oscilloscope* simple, but at the same time, as realistic as possible – so, the aim is to clean up the layout of the screen and add more, genuinely useful functions. I've changed the layout of the screen to be more symmetrical, there are now eight squares along the horizontal and four in the vertical, giving a nice even 32 squares. I've changed the maximum voltage to 4V, so we can always see the top of the waveform and to keep it nice and even. I've also adjusted the ADC values in the code to match 3V on the screen. This can all be seen in the code available on *EPE's* website.

So, the features we are going to add are:

1) Ability to measure voltage level on screen
2) Rising-edge/falling-edge triggers
3) Automatic sampling/normal sampling
4) Stop sampling and HOLD current sample on screen
5) Infinite persistence

Buttons first

Let's get started, in order to change values on the fly without having to program our LPLC board every time, we're going to need a few buttons – four is sufficient. This allows us to cycle through various options. One button can be used solely for stopping the sampling, and holding the current trace on the screen. The next button can be used to select the various items around the screen; we can change each item to yellow to indicate it is selected. The next two buttons are 'up' and 'down' for the current selected item.

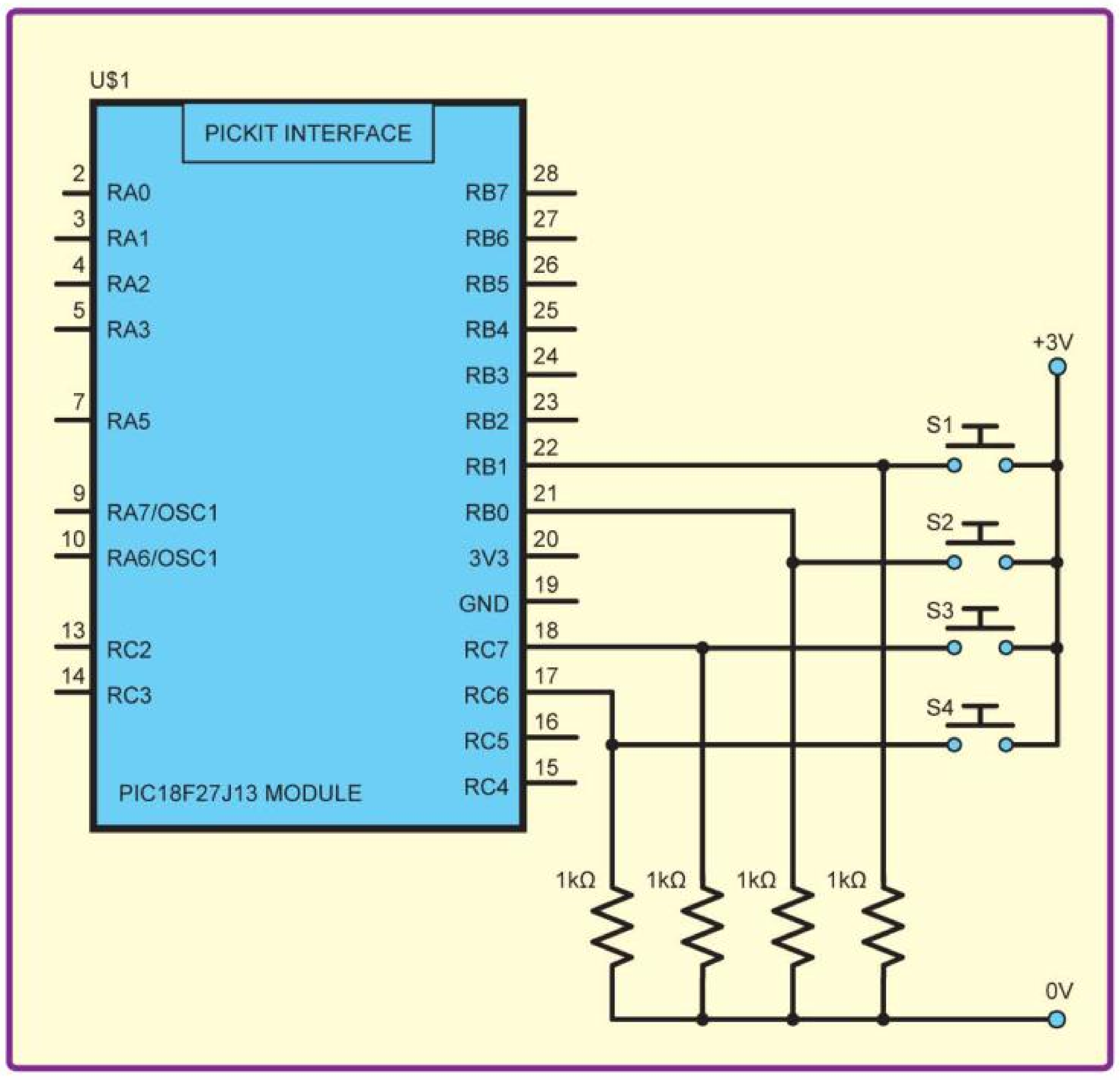


Fig.1. Push-button schematic

Now we must connect some push buttons. Ideally, we want to have our inputs on the PIC toggle between high and low to indicate the button has been pressed. The best way is to connect each of our inputs to ground using 1kΩ resistors. Then we connect one side of our push buttons to the input pin of the PIC and the other side of the button to our 3V supply. We expect logic 0 at all times – *except* when a button is pressed and then we see logic 1.

Now we have the hardware sorted out, we need some code to act on the button presses. For our input pins for the buttons, we're going to use the following ports on the PIC: PORTC6, PORTC7, PORTB0 and PORTB1. We initialise these ports as inputs in our existing function `HardwareInit()` in the **hardware.c** file.

```
TRISBbits.TRISB0 = 1;       // Setup as Input
TRISBbits.TRISB1 = 1;       // Setup as Input
TRISCbits.TRISC6 = 1;       // Setup as Input
TRISCbits.TRISC7 = 1;       // Setup as Input
```

Next, we create a function called `CheckforButtonPress()` in the **hardware.c** file (Listing 1), which checks to see which button has been pressed; it will either toggle between the options or increase/decrease a value. PORTC6 will be used to halt or stop the current sampling. PORTC7 will be used to select between the various on-screen options. PORTB0 and PORTB1 will increase and decrease the selected on-screen value. Only the code for 'HOLD' and 'SELECT' are shown – the rest of the code can be downloaded from the *EPE* website. We add the `delayms(500);` command as a debounce for our buttons, so that they are not pressed more than once in a given (short) period.

Listing 1

```
void CheckforButtonPress(void) {
        if (PORTCbits.RC6) { // Hold Sampling Button
                // Assert HOLD_SAMPLING if not already asserted
                // and stop writing to screen
                // Clear Screen and start sampling again if
                // HOLD_SAMPLING is already asserted
                if (HOLD_SAMPLING) {
                        LCD_Clear(BLACK);
                        POINT_COLOR = GRAY;
                        LCD_DrawRectangle(10,10,239,319);
                        LCD_DrawScreenWords(BUTTON_SELECT);
                        HOLD_SAMPLING = 0;
                } else {
                        HOLD_SAMPLING = 1;
                }
                delayms(500);
        }


        // Select Screen words code
        if (PORTCbits.RC7) {
                // Checks Button Select
                // Increment button selected, LCDDrawScreenWords
                // function chooses colour of on screen selection
                if (BUTTON_SELECT == 3) {
                        BUTTON_SELECT = 0;
                } else {
                        BUTTON_SELECT++;
                }
                LCD_DrawScreenWords(BUTTON_SELECT);
                delayms(500);
                }


                // Check if Up Button has been pressed and increment
                // values if needed and debounce button

                // Check if Down Button has been pressed and decrement

                // values if needed and debounce button
}
```

Listing 2

```
void LCD_DrawTriggerLine(void) {
        // Initialise variables for function, calculate mV line
        unsigned char trig_char[16];
        unsigned long int trig_temp = TRIGGER_LINE-11;
        trig_temp = trig_temp * 190;
        trig_temp = trig_temp / 10;
        sprintf(trig_char, “%d mV”, trig_temp);

        // Check if TRIGGER_LINE has been selected,
        // if selected colour is yellow, red if not
        // Write our new line to the screen
}
```

Trigger level

Now we look at what we want on the screen. We'll need an on-screen trigger level. We create a function `LCD_DrawTriggerLine()` and a global variable TRIGGER_LINE to draw the line on the screen. This function will be called every time a new capture is available. Now all we have to do is increment/decrement the TRIGGER_LINE variable and the change will be seen next time a capture is available. Listing 2 (below) is the code for the function. In the function, we must remove whatever line we had previously. I've also added the trigger-level value to the bottom-right of the screen in mV. So, now we can set and see the trigger-level value.

It's also nice to have a measurement line, which will indicate the voltage level where this line sits. This is just the same code as in Listing 2, except we use the MEASURE_LINE variable instead of TRIGGER_LINE.

Rising/falling-edge triggers

Now we want to see a transition from high to low or low to high (also known as rising and falling edges). This is one of the most useful aspects of any oscilloscope, as it allows us to see if a signal is behaving as it should. From the code above, we now have a TRIGGER_LINE variable, which we can use to see if a value has gone from high to low.

Lets create a variable called EDGE_SAMPLE, which indicates we want a rising edge when it is 1 and we want a falling edge when it is 0.

Now, in order to find a rising or falling edge, we need to look at the ADC values we're capturing. We edit the function `handle_int_ADC` in `lplc_adc.c` and add the following lines inside the `if(!samples_ready)` statement. What we want to do here is look at the previous ADC value, compare it against the new ADC value and see if it is either a rising or falling edge. If either are seen, we assert the RISING_EDGE and FALLING_EDGE global variables – see Listing 3. We will come back to this later to bring it all together.

Automatic/normal sampling

Another useful feature of the oscilloscope is the Automatic or Normal sampling features. Automatic is just a free running sampling mode, which updates the screen every time a new sample is ready. However, the Normal sampling feature is much more useful, in that

Listing 3

```
// Take the last sample value and store it in prev_sample,
// Increment the sample_index to store new adc value
prev_sample = samples[current_buff][sample_index];
samples[current_buff][sample_index++] = 11 + ((current_adc << 1)/13);
// scale the sample to our range of 11-238, or 11 + 0-227,
// Take latest sample and store in next_sample
next_sample = samples[current_buff][sample_index];

// Compare previous sample and current sample versus our
// TRIGGER_LINE global variable
if((prev_sample <= (TRIGGER_LINE)) && (next_sample >= TRIGGER_LINE) && (!RISING_EDGE)) {
        FALLING_EDGE = 1;
}
if((prev_sample >= (TRIGGER_LINE)) && (next_sample <= TRIGGER_LINE) && (!FALLING_EDGE)) {
        RISING_EDGE = 1;
}
```

we only want to update the screen when a trigger event has happened. Something like our RISING_EDGE and FALLING_EDGE triggers mentioned earlier.

Ideally, we would like our transitioning signal to occur in the middle of the screen, but I think we will just stick to triggering anywhere on the screen. We need some Boolean algebra here – we want to enter this `if` statement if the following conditions are true:

1) It is not the first sample
2) AND auto sampling is not enabled
3) OR a falling edge has been seen
   when EDGE_SAMPLE is a logic 0
4) OR a rising edge has been seen when
   EDGE_SAMPLE is a logic 1

```
  if  ((!first_sample)  &&  ((!AUTO_SAMPLE)  ||
((FALLING_EDGE  &&  !EDGE_SAMPLE)  ||  (RISING_
EDGE && EDGE_SAMPLE)))) {
        ---
        }
```

Stop Sampling and HOLD current sample on screen

Suppose we want to capture a signal doing something on the screen, but we keep losing it. What we need is the ability to press 'stop' on a signal we've just captured and have a look at it. Using a unique button dedicated to holding a signal, we can easily capture our on-screen signal. First, we grab ourselves a global variable, `HOLD_SAMPLING`. When it is 0, we let the sampling run free under whatever conditions it's in. When it is a 1, we stop writing to the screen. We don't necessarily have to stop the ADC from running (we could if we wanted to and thus save some power). We can implement this change in the exact same place as earlier when we change from automatic to normal sampling. We just add a Boolean AND to our `if` statement and we're good to go. When we press the button again, the hold is released and we continue writing to the screen, see below:

```
  if  ((!first_sample)  &&  (HOLD_SAMPLING)  &&
((!AUTO_SAMPLE)  ||  ((FALLING_EDGE  &&  !EDGE_
SAMPLE)  ||  (RISING_EDGE && EDGE_SAMPLE)))) {

        ---

  }
```

I also thought it would be nice to remark on the fact that we are not running, by printing the word 'HOLD' in big bold red letters, top right of the screen.

Infinite Persistence

Infinite persistence is another useful feature. It's where the sample line persists on screen or doesn't go away. This is actually very easy to do. At the moment, we draw over our last samples in black and draw our new samples in green (in effect deleting our last line). Instead of deleting our previous lines, we can just leave them there and draw our new samples wherever they may lie. In the main loop, when we write over our existing green line, we add a simple `if` statement and we have our infinite persistence.

When is infinite persistence actually useful? Imagine we want to see how a voltage rail into our device behaves? Maybe we've added a Wi-Fi or GSM module to our device. Both of these technologies cause regular instantaneous 2A current draws when trying to connect. If there isn't sufficient current or it doesn't provide the current fast enough, then we will get a voltage drop. This is when infinite persistence becomes particularly useful. We can capture our voltage minimums, maximums and any spurious voltage spikes.

```
  if(!INFINITE_PERSISTENCE) {
        POINT_COLOR = BLACK;
        LCD_DrawLine(old_samples[lpy-1],
10+lpy, old_samples[lpy], 11+lpy );
  }
```

Wrapping it all up

So, we've connected buttons to allow us to interface with our board – plus, we've written the code to allow us do something with those buttons. We've added edge-sample triggering for free running (AUTO sampling) and triggered running (NORMAL sampling). We've also added a stop button and infinite persistence. We bring it altogether and now we can look at various signal inputs connected to our PIC. Fig.2 shows a poor rise time from the output of a function generator. Alternatively, we could output a pulse-width modulation signal from one our of CCP (Capture/Compare/PWM) pins and connect it up to our input PORTA0.

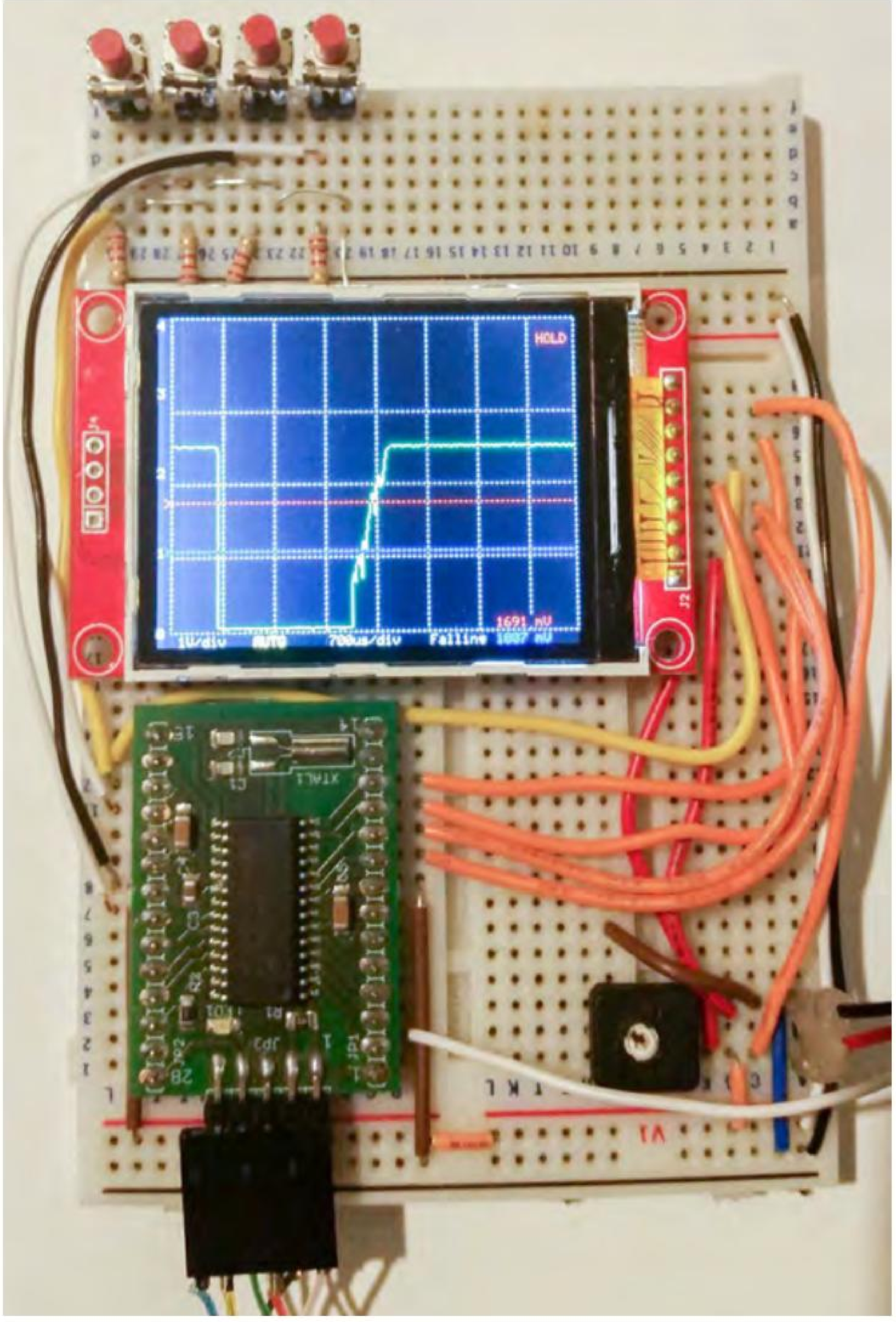


Fig.2. LPLC Oscilloscope *showing a stopped rising-edge trigger*

To improve on this design, there are a few things we could add. One of the inherent problems with the current PIC is we assume we know the value and consistency of the PIC's VDD input voltage – after all, this is what our ADC values are based on. If this changes, then our measurements will be incorrect. In order to measure our own voltage without using any external measuring circuits, we could use the band gap reference that other PIC microcontrollers use (like the PIC24F family). This is a stable, internally regulated voltage reference at 1.2V. We could use this reference to measure input voltages.

We could also add improved circuit protection against ESD and damaging voltage spikes. We will come back to our DIY *LPLC Oscilloscope* design later in the year with a bigger and better PIC and use it to increase the sample rate into the MHz range.

Next month

In November's issue we'll take a look at MikroElektronika's 'mikroMedia for PIC32'. This board uses the same size screen as our *LPLC Oscilloscope*, but includes a touch screen interface. It uses the PIC32MX460F512L and a stereo MP3 codec chip that allows us to develop multimedia apps with ease.

Not all of Mike's technology tinkering and discussion makes it to print. You can follow the rest of it on Twitter at @MikePOKeeffe, and from his blog at mikepokeeffe.blogspot.com

VAT
things
wirelessthings.net/epe-offer
@wirelessthings

AUDIO OUT

By Jake Rothman

RIAA equalisation – Part 4

Testing RIAA amplifiers

A convenient way to test an RIAA input is to pass a signal from a signal generator through an inverse RIAA network. This has the opposite frequency response of an RIAA pre-amplifier's equalisation curve, which hopefully results in an overall flat response when combined. If it deviates, it indicates a problem somewhere, such as bad design or poor component tolerance. A suitable test set-up is given in Fig.28. Fig.29 shows an inverse RIAA circuit based on a design by Reg Williamson, then modified by Lipshitz and Jung in January 1980 in *The Audio Amateur* – followed by a few tweaks of my own. It's a passive network, with high 'insertion loss' (loading) effect, but this is advantageous since RIAA inputs have very high gain. This enables a better match to the (relatively) high voltage output of signal generators. The curve of the inverse RIAA filter is in Fig.30. The circuit can be built up onto a phono plug, as shown in Fig.31. The whole circuit can then be plugged into left and right inputs in turn to check channel matching.

Amplifier check

I was recently asked to look over a Sansui A-80 amplifier brought into the workshop with the fault, 'sounds louder on one side'. I checked the phono stage RIAA response, which gave a normal curve shown in Fig.32. In conjunction with the inverse RIAA filter a minor deviation of +0.5dB broadly centred on 2kHz was observed, shown in Fig.33. The 'fault' turned out to be a 30dB 4kHz dip in one of the septuagenarian owner's ears. The small 2kHz rise was enhancing, filling the slight droop most cartridges exhibit between the inductive droop and stylus resonance. Was it deliberate 'voicing' on the designer's part?

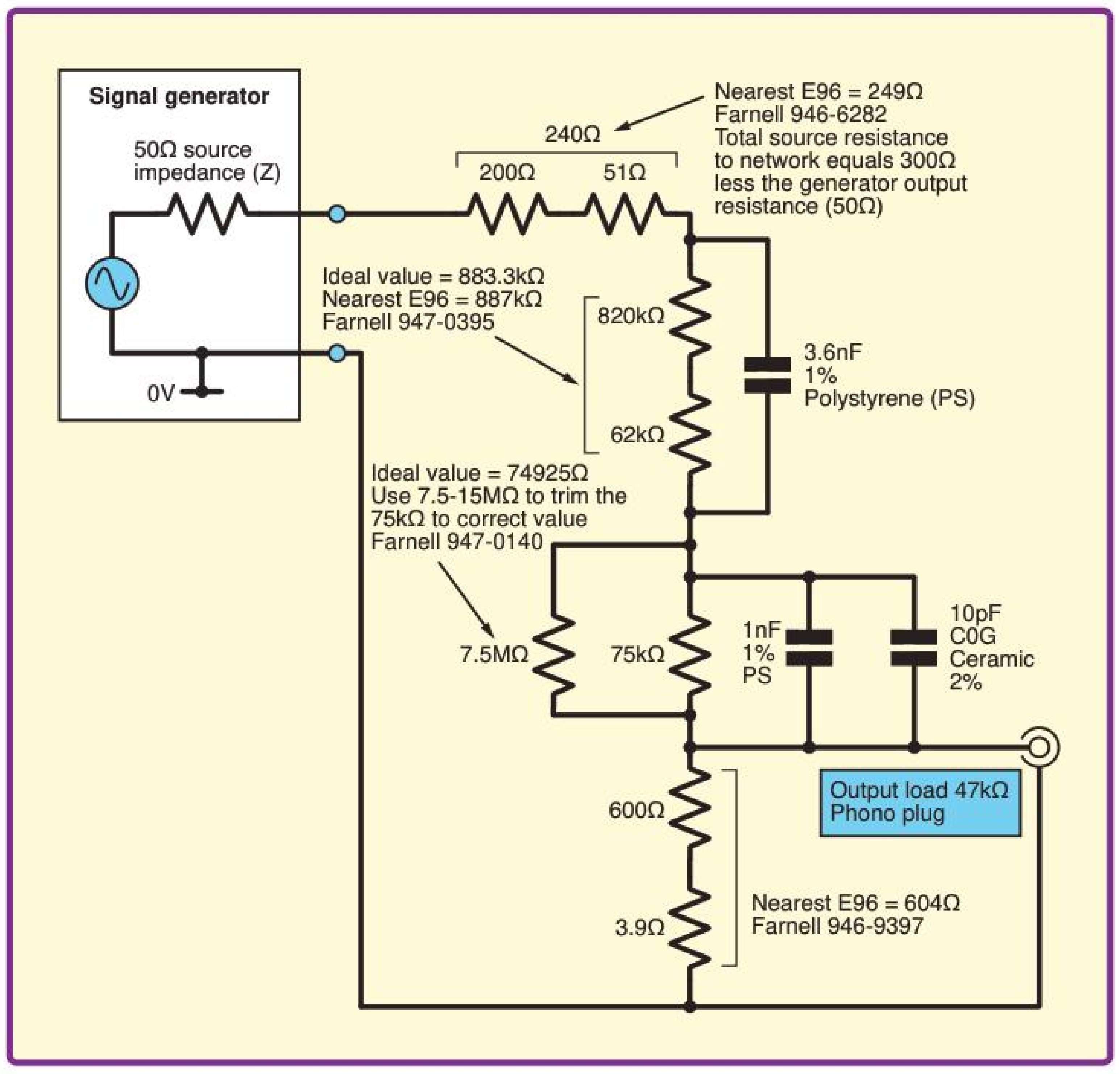

Signal generator
50Ω source impedance (Z)
0V
240Ω
200Ω
51Ω
Nearest E96 = 249Ω
Farnell 946-6282
Total source resistance to network equals 300Ω less the generator output resistance (50Ω)
Ideal value = 883.3kΩ
Nearest E96 = 887kΩ
Farnell 947-0395
820kΩ
62kΩ
3.6nF
1%
Polystyrene (PS)
Ideal value = 74925Ω
Use 7.5-15MΩ to trim the 75kΩ to correct value
Farnell 947-0140
7.5MΩ
75kΩ
1nF
1%
PS
10pF
C0G
Ceramic
2%
Output load 47kΩ
Phono plug
600Ω
3.9Ω
Nearest E96 = 604Ω
Farnell 946-9397

Fig.29. Inverse RIAA network – it uses some funny resistor values. The Farnell 1% MSR25 series resistors are recommended

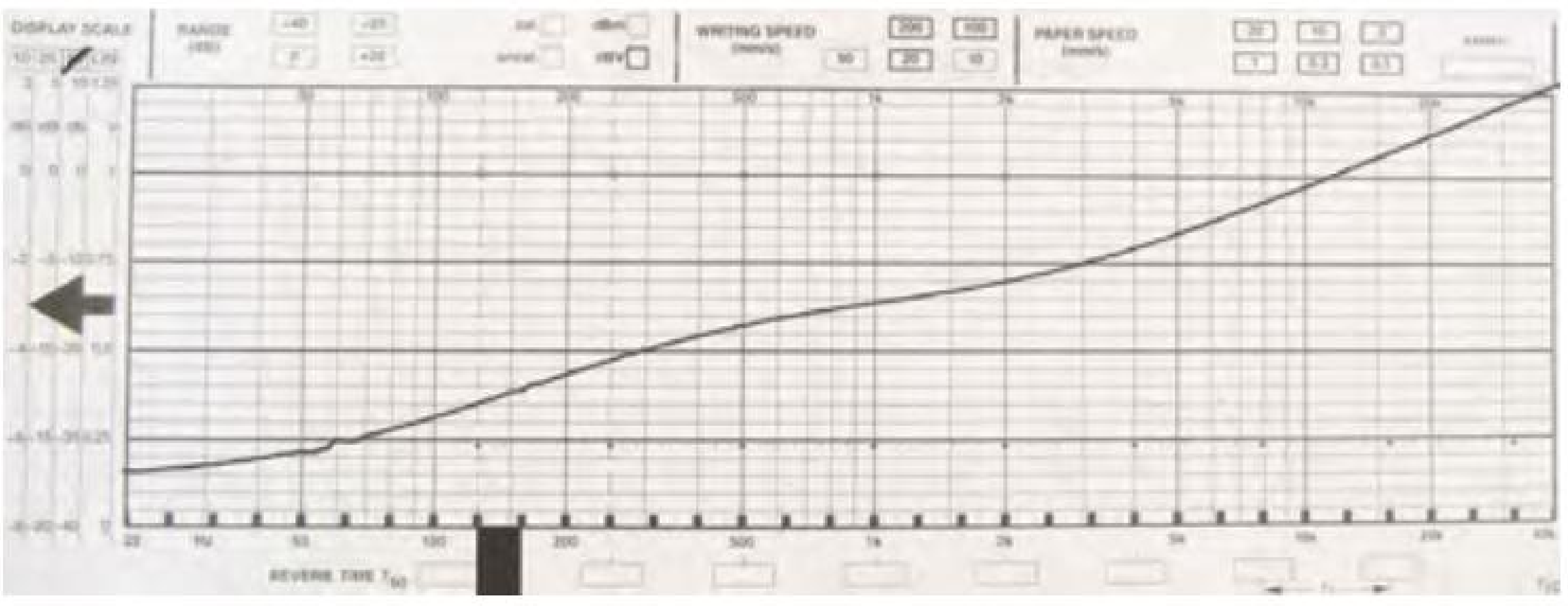


Fig.30. Inverse RIAA network curve

System synergy

I checked the Sansui amplifier's moving coil (MC) input with an Audio-Technica AT32E cartridge, which sounded hissy with high-frequency loss. The MC input stage was simply a common-emitter circuit tacked onto the moving-magnet input. I then tried using audio-matching transformers via the moving-magnet input and it sounded fantastic! An accidental synergy had occurred, a serendipitous event all audio system designers experience sometimes. The hiss was subjectively lower than with a 5534 RIAA input using the transformers. The noise spectrum was also biased more towards the extreme high frequencies, making it much less noticeable. The transformers used were standard industrial audio transformers from Vigortronic (Fig.36) costing about £9.00 each from Rapid Electronics – *not* the crazy expensive 'Hi-Fi' variety. Investigation found the Sansui's RIAA stage had a JFET on the input. Such JFET designs usually give lower noise with high source impedances compared with bipolar transistor versions. It appeared this was the case here, with the inductive load of the moving-coil cartridge reflected through the transformer accidently providing excellent impedance matching. As every electronic circuit designer knows, when two circuit blocks connected together give unexpectedly good results, a new design is in the making.

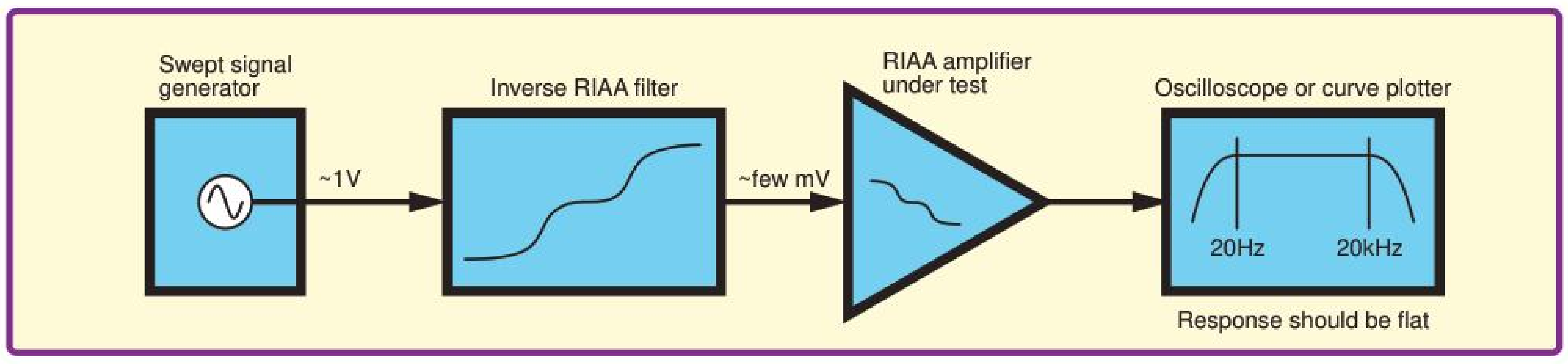

Swept signal generator
~1V
Inverse RIAA filter
~few mV
RIAA amplifier under test
Oscilloscope or curve plotter
20Hz
20kHz
Response should be flat

Fig.28. Using an inverse RIAA network to test RIAA inputs – the output should be flat

The circuit of the Sansui input stage is given in Fig.35, with my transformer connections shown in Fig.37. It is a fairly simple four-transistor arrangement, similar to many small power amplifiers. The standard 'ring-of-three' feedback system, comprising input voltage to current (transconductance) stage, voltage amplifier stage (VAS),

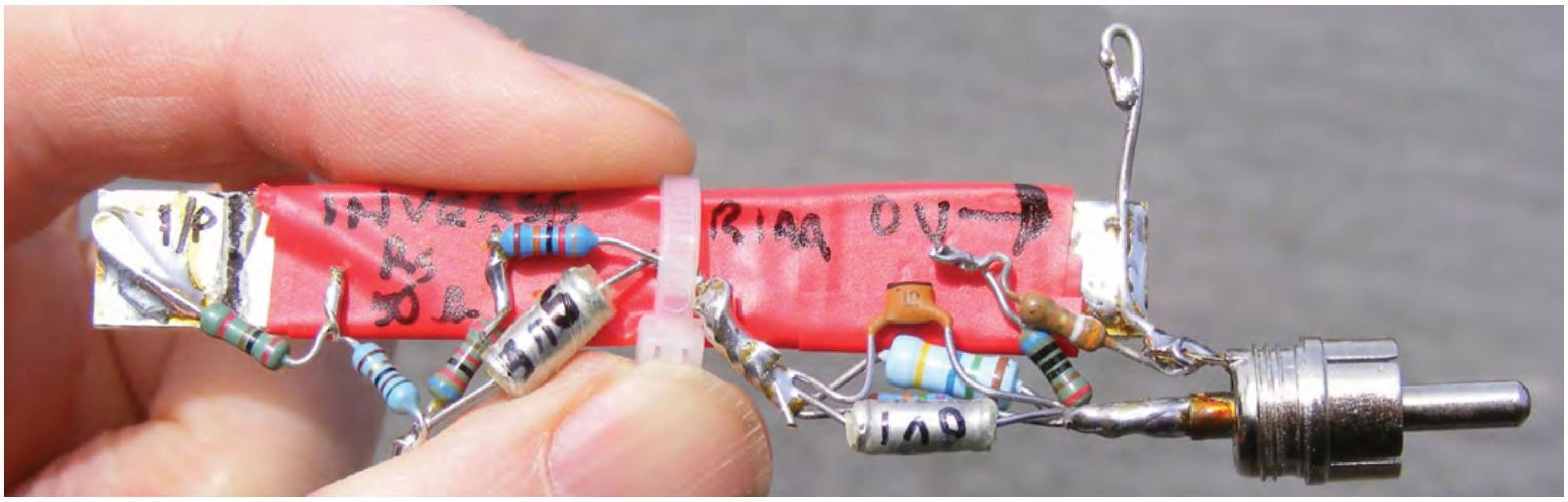

I/P
INVERSE RIAA
RIAA 0V

Fig.31. Photo of 'hard-wired' inverse RIAA network – it plugs into an RIAA amplifier input

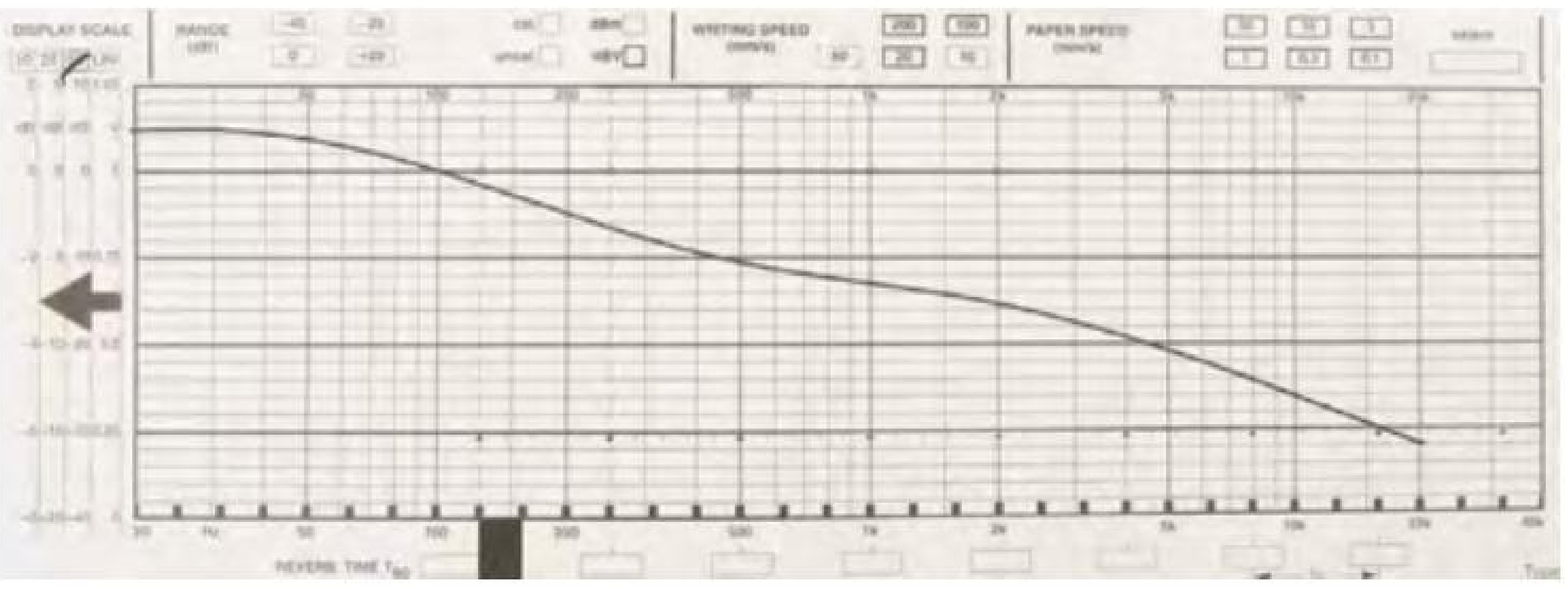


Fig.32. Sansui A80 RIAA response curve

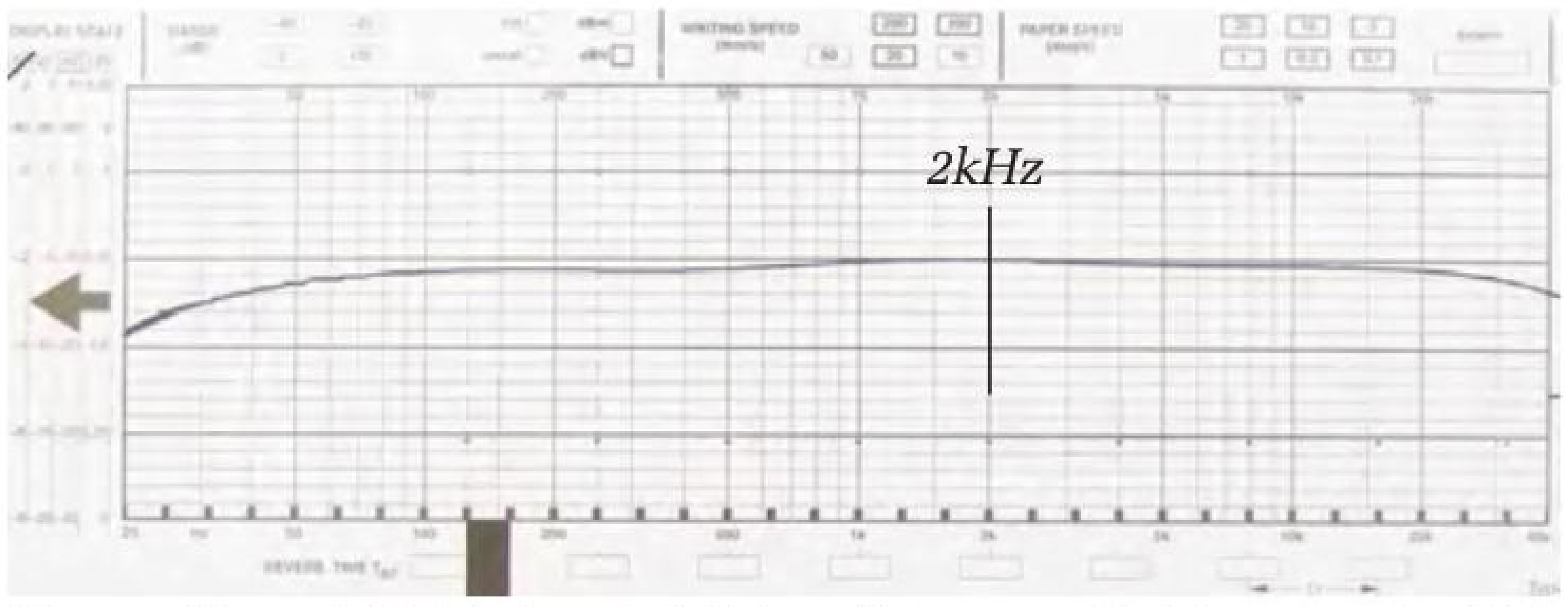


Fig.33. Sansui RIAA input fed by all inverse RIAA network. Note slight response bump at 2kHz

Fig.34. Sansui A-80 amplifier

common-emitter stage and output stage. There are some variations to the standard theme, the input transistor is a JFET, the VAS stage is 'upside down' using a PNP device, enabling a more common N-channel JFET to be used. Finally, the output stage is a class-AB push-pull, emitter-follower design. This will not generate crossover distortion on normal signal levels, but will have the current capacity to deal with nasty scratches without overloading by going momentarily into class B. This will sound less objectionable than a hard-clipped class A.

FET frenzy

Unfortunately, as with most Japanese audio circuits, the Sansui used unusual semiconductors. The 2SK163, (marked K163) JFET from NEC was difficult to source. The other devices were all easily substituted. The K163 and other rare audio JFETs are available as NOS 'new old stock' from Donberg Electronics in Ireland. I measured the transconductance (g_{fs}) of the K163 with the Peak DCA75 and found it to be in accordance with the data sheet at 17mA/V. This is much higher than standard JFETs, such as the 2N5457, which typically achieve 3-4mA/V. This would reduce the open-loop gain of the circuit by around four times and increase the noise level, since the second stage would have to do more work. The

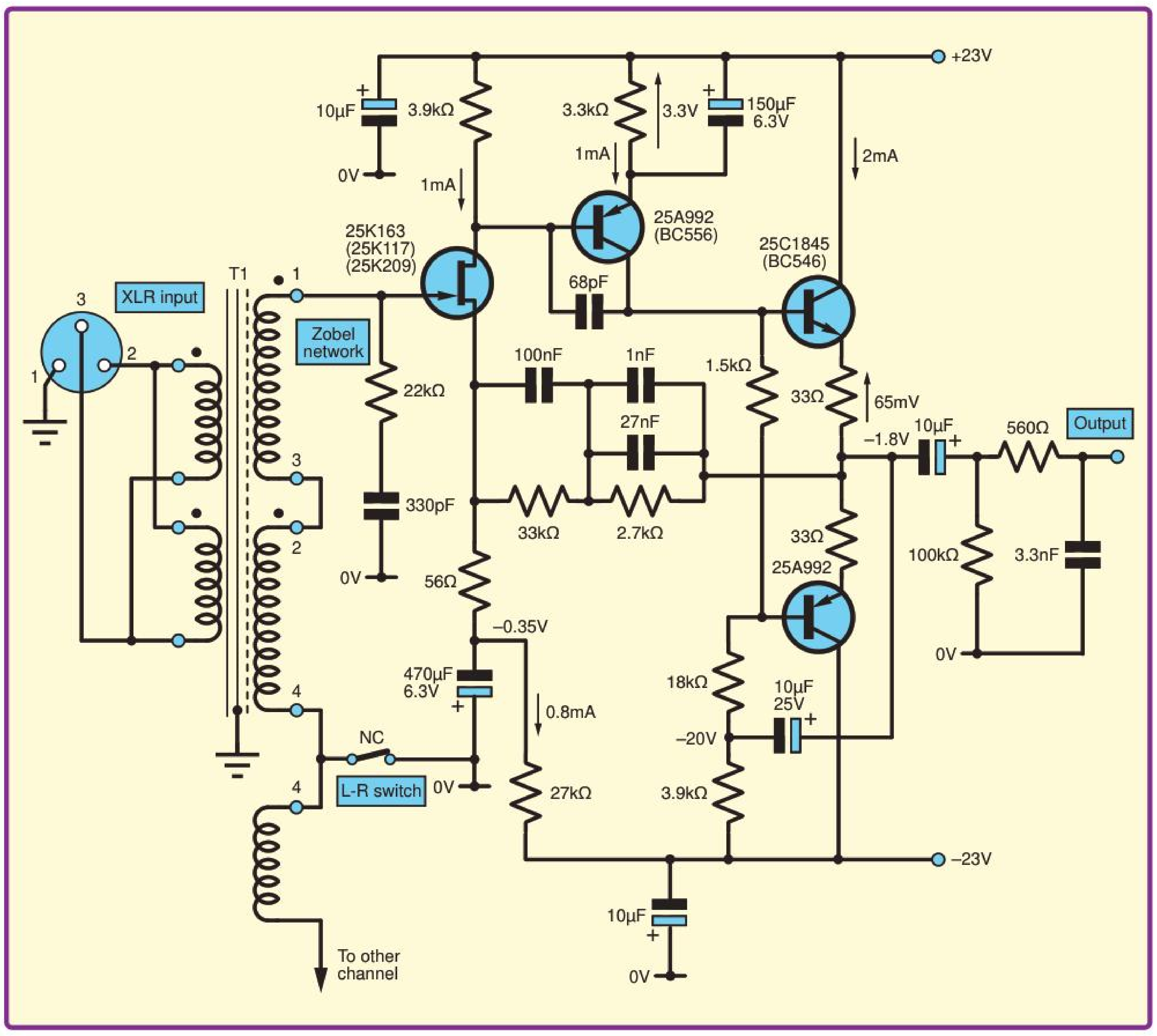

+23V
10µF
3.9kΩ
3.3kΩ
3.3V
150µF 6.3V
0V
1mA
1mA
2mA
25A992 (BC556)
25K163 (25K117) (25K209)
25C1845 (BC546)
T1
1
3
XLR input
2
1
68pF
Zobel network
100nF
1nF
1.5kΩ
22kΩ
27nF
33Ω
65mV
–1.8V
10µF
560Ω
Output
3
330pF
33kΩ
2.7kΩ
33Ω
2
25A992
100kΩ
3.3nF
0V
56Ω
–0.35V
0V
470µF 6.3V
4
0.8mA
18kΩ
10µF 25V
NC
–20V
4
L-R switch
0V
27kΩ
3.9kΩ
–23V
10µF
To other channel
0V

Fig.35. Sansui A-80 RIAA input stage – a simple discrete JFET input design, with a few minor changes and transformers added. The switch on the transformer ground allows the left minus right signal to be listened to, which allows the anti-skating force on the pickup to be set to optimum by minimising distortion on vocal sibilants

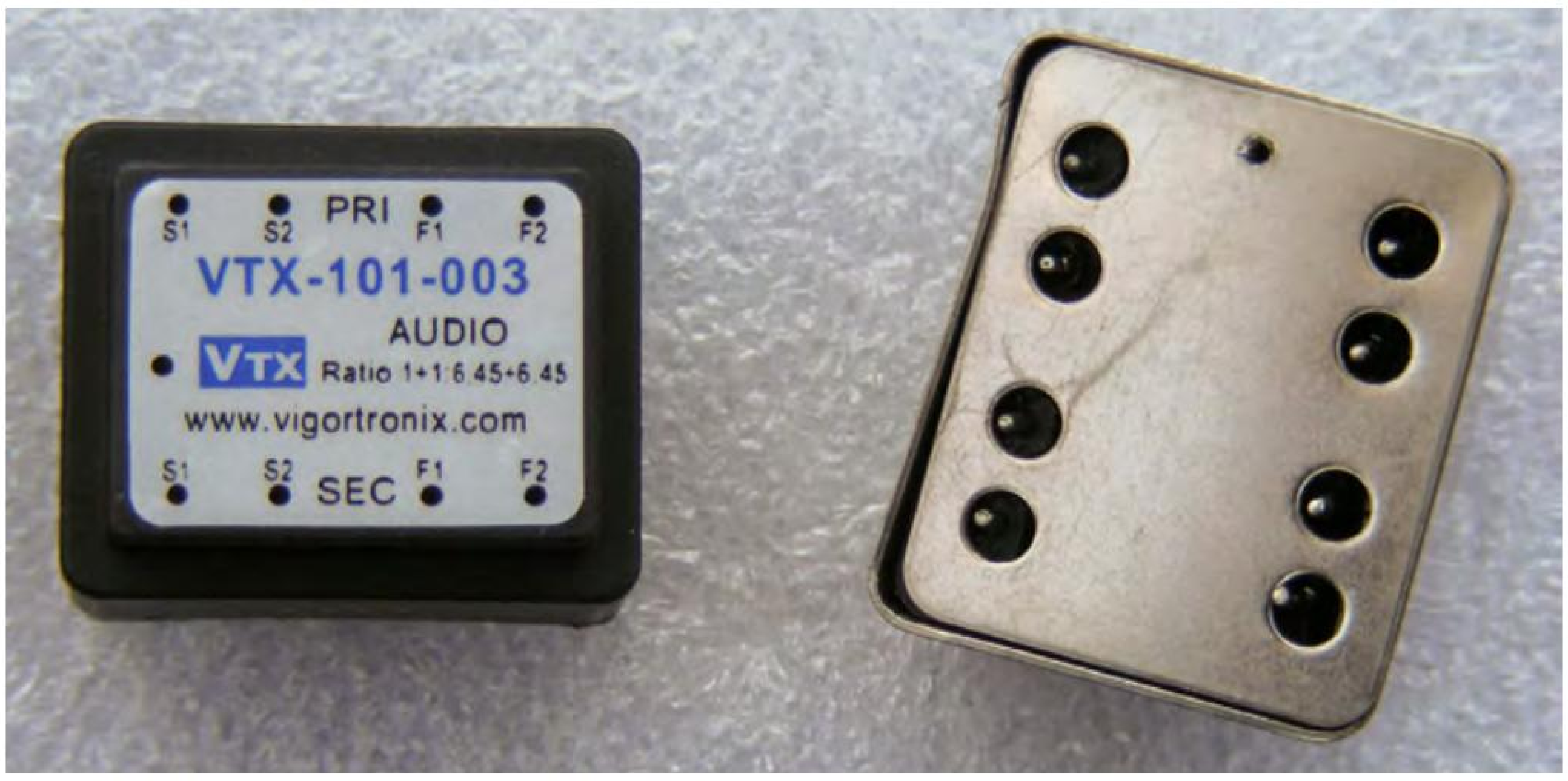

S1
S2
PRI
F1
F2
VTX-101-003
AUDIO
VTX
Ratio 1+1:6.45+6.45
www.vigortronix.com
S1
S2
SEC
F1
F2

Fig.36. Vigortronix 003 transformers used for converting moving-magnet to moving-coil inputs. Note the essential mumetal screening can to prevent hum pick-up

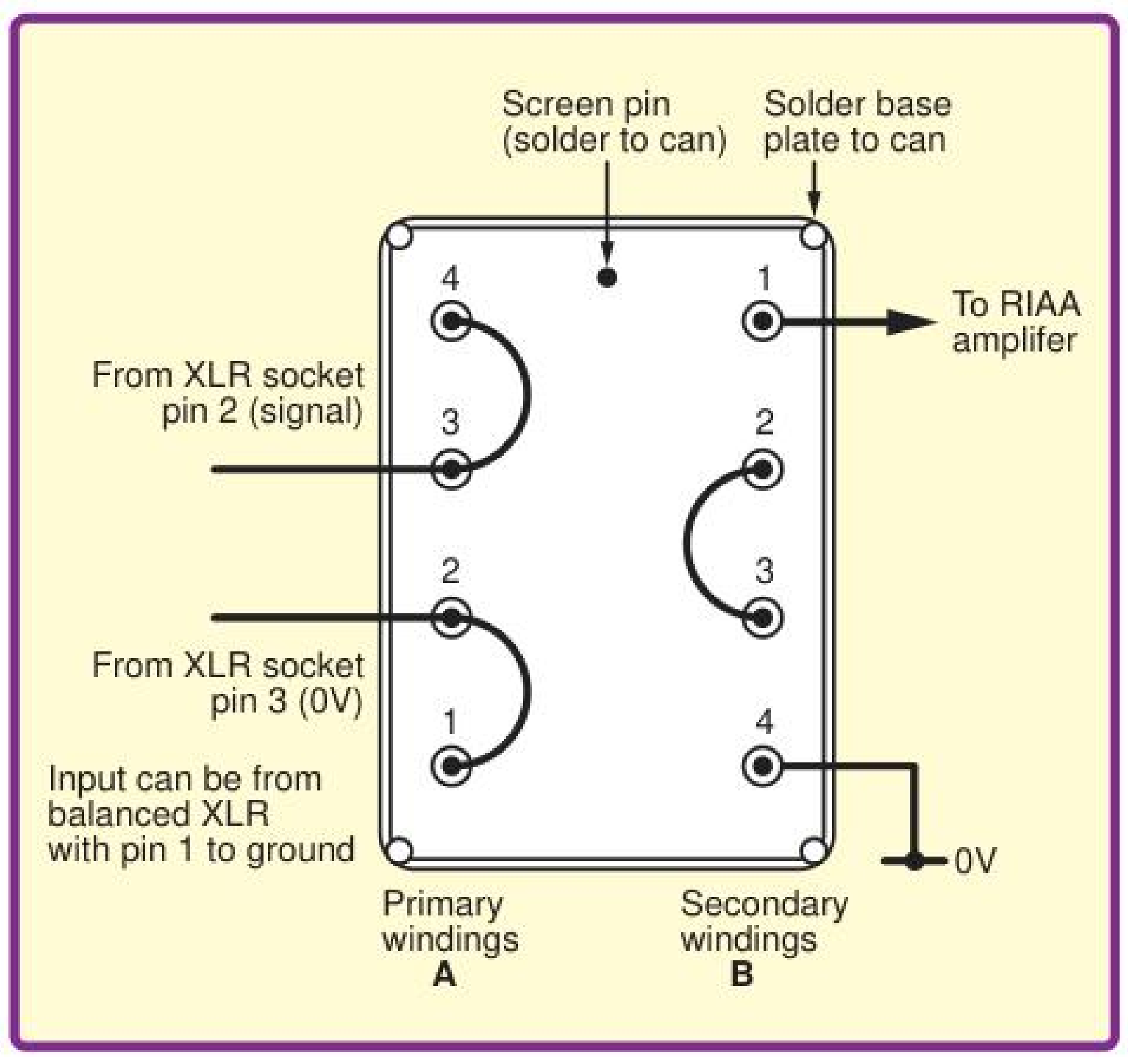

Screen pin (solder to can)
Solder base plate to can
4
1
To RIAA amplifer
From XLR socket pin 2 (signal)
3
2
2
3
From XLR socket pin 3 (0V)
1
4
Input can be from balanced XLR with pin 1 to ground
0V
Primary windings
A
Secondary windings
B

Fig.37. Wiring for a Vigortronix VTX-101-003 moving-coil step-up transformer (pin view)

K163 is one of those rare 1980s audio JFETs that uses a finely interdigitated channel (see Fig.38) to achieve the high transconductance and low noise. Other suitable types are the 2SK117 and 2SK170. If a P-channel JFET is used, such as the 2SJ74, the circuit could be made to work on a single +48V rail. One side effect is high capacitance, but this is not a problem in this application. Another advantage of FETs is that no input capacitor is needed in this circuit.

Toshiba still make some audio FETs, albeit only in surface mount packs, such as the 2SK209-Y which is designed for condenser microphone head amps. These cost around 25p from Mouser UK and meet all the specs. I expect to pay commodity prices for my parts and I have no truck with the 'antique-dealer' pricing endemic in the audio parts industry. The K163 has an unusually high maximum channel voltage of 50V, but the device is only exposed to 17.5V if the rails are ramped up slowly. This allows 30V TO92 package devices to be used. For those interested in all things JFET, Erno Borbely is the circuit designer to watch.

Wind up

Finally, to end on a 'sad' note, a friend of mine gave me his defunct £500 moving-coil cartridge that had only lasted for ten plays. In an attempt to prevent 'extraneous resonance' from a casing, the designer had left the coils exposed – typical Hi-Fi madness. The inevitable happened, one of the fine wires got caught and broke. In a failed attempt to repair the unit, I was able to take a photos showing the inner workings of a moving-coil in Fig.39, and just for interest, there is a ceramic cartridge internal structure shown in Fig.40.

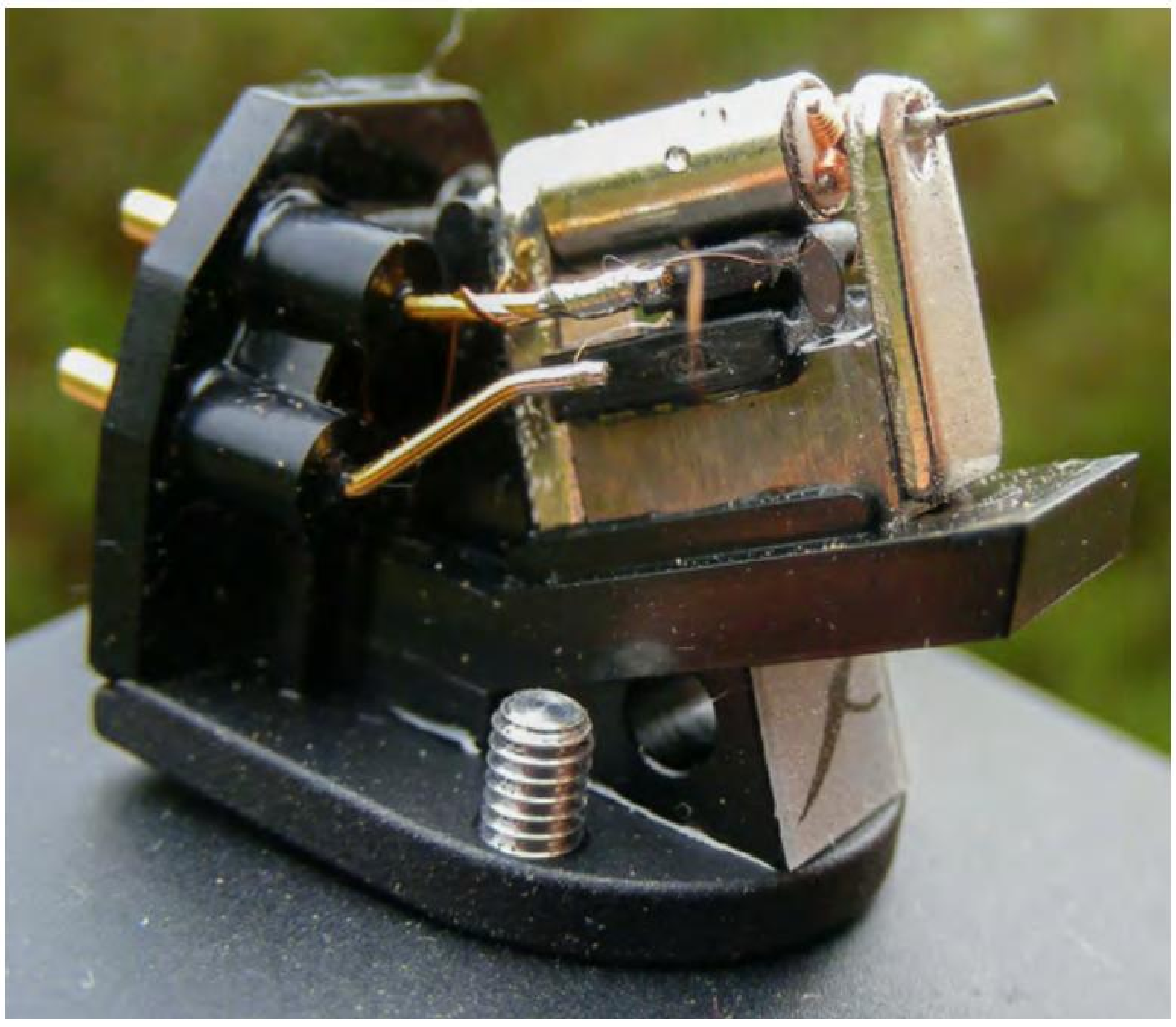


Fig.39. Sumiko Blackbird moving-coil cartridge with broken wire. The 4-coil armature can be seen

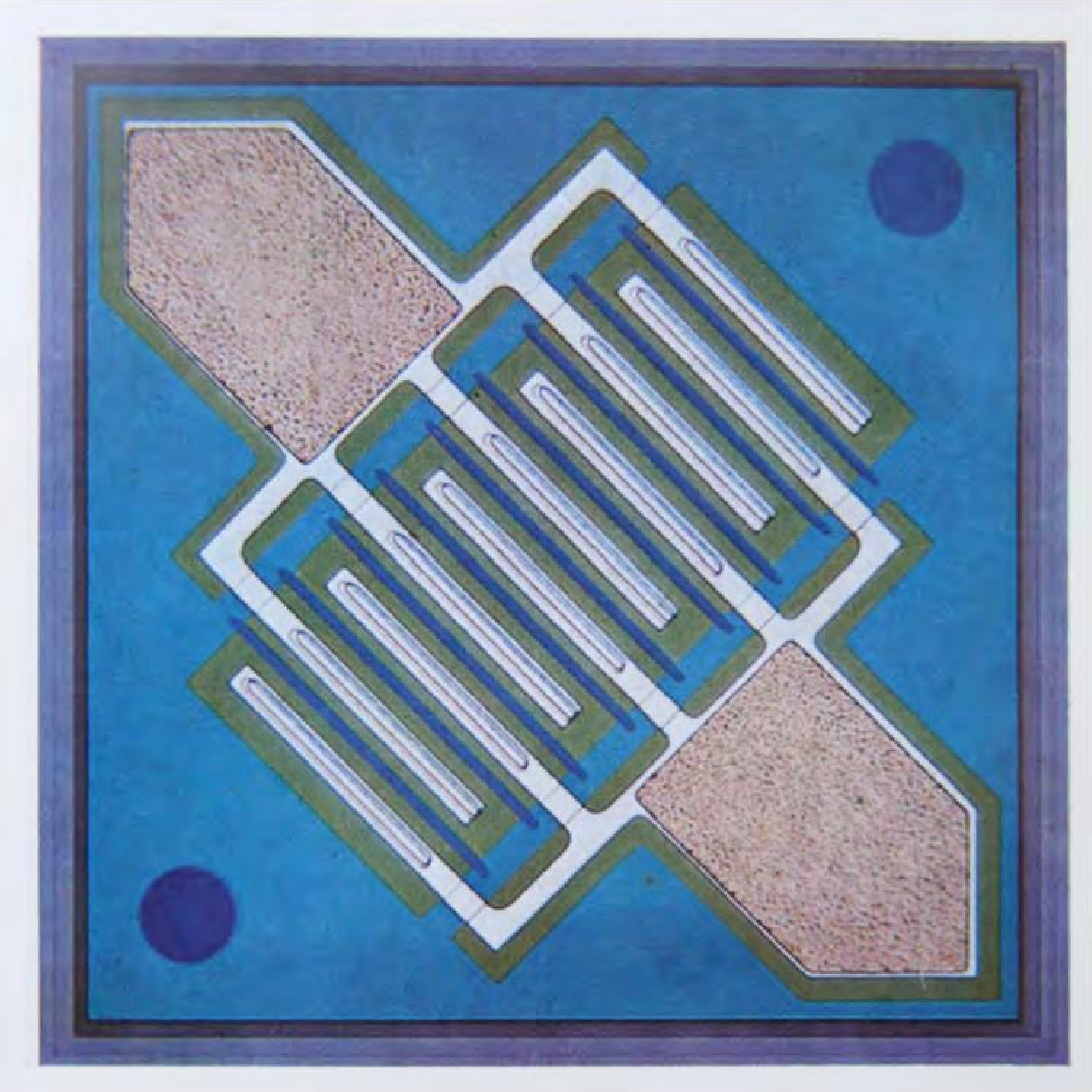


Fig.38. An example of an 'interdigitated' JFET – an early low-noise audio design, the BFW10 – from Field-Effect Transisors, *Mullard Ltd, 1972; Plate A p13*

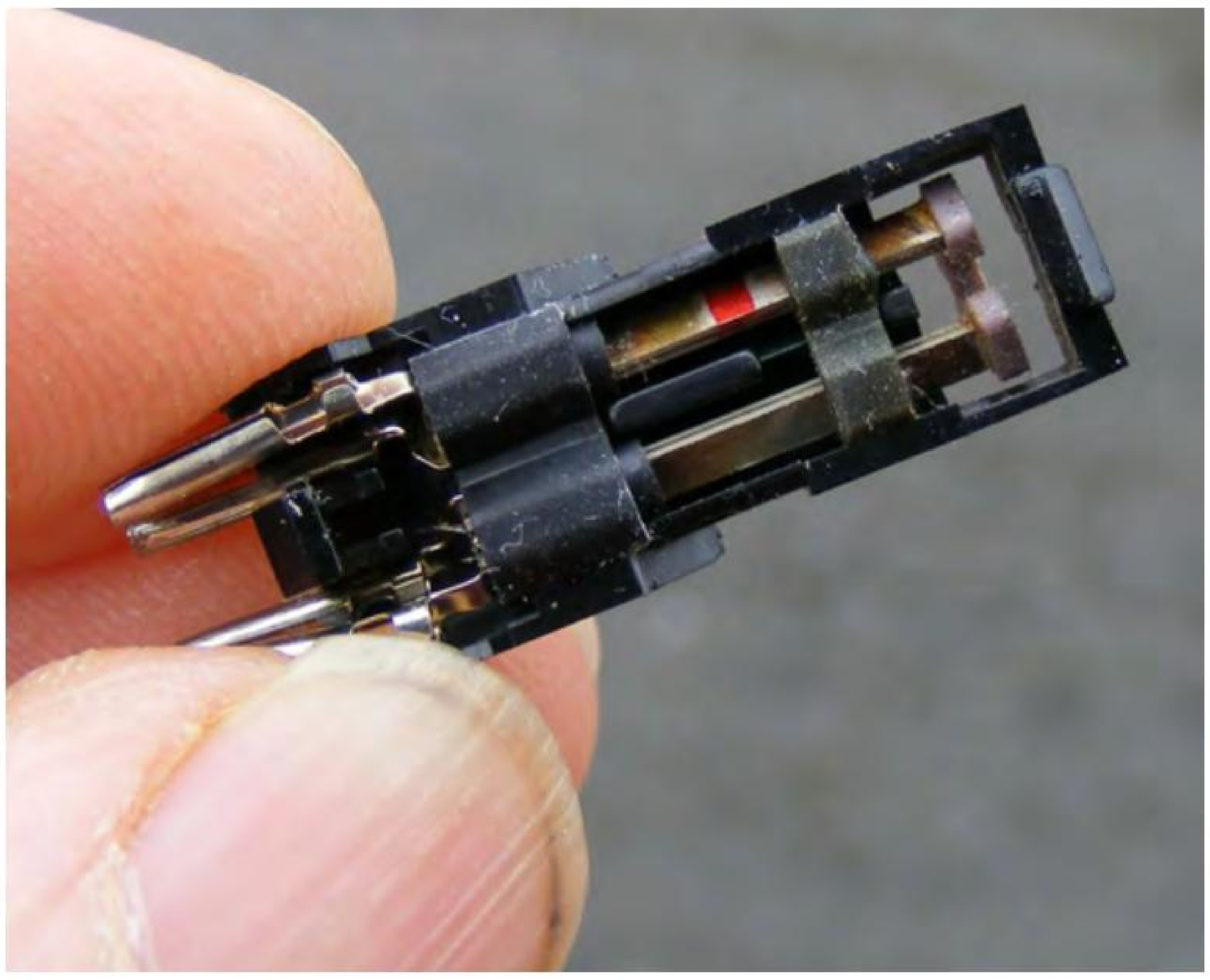
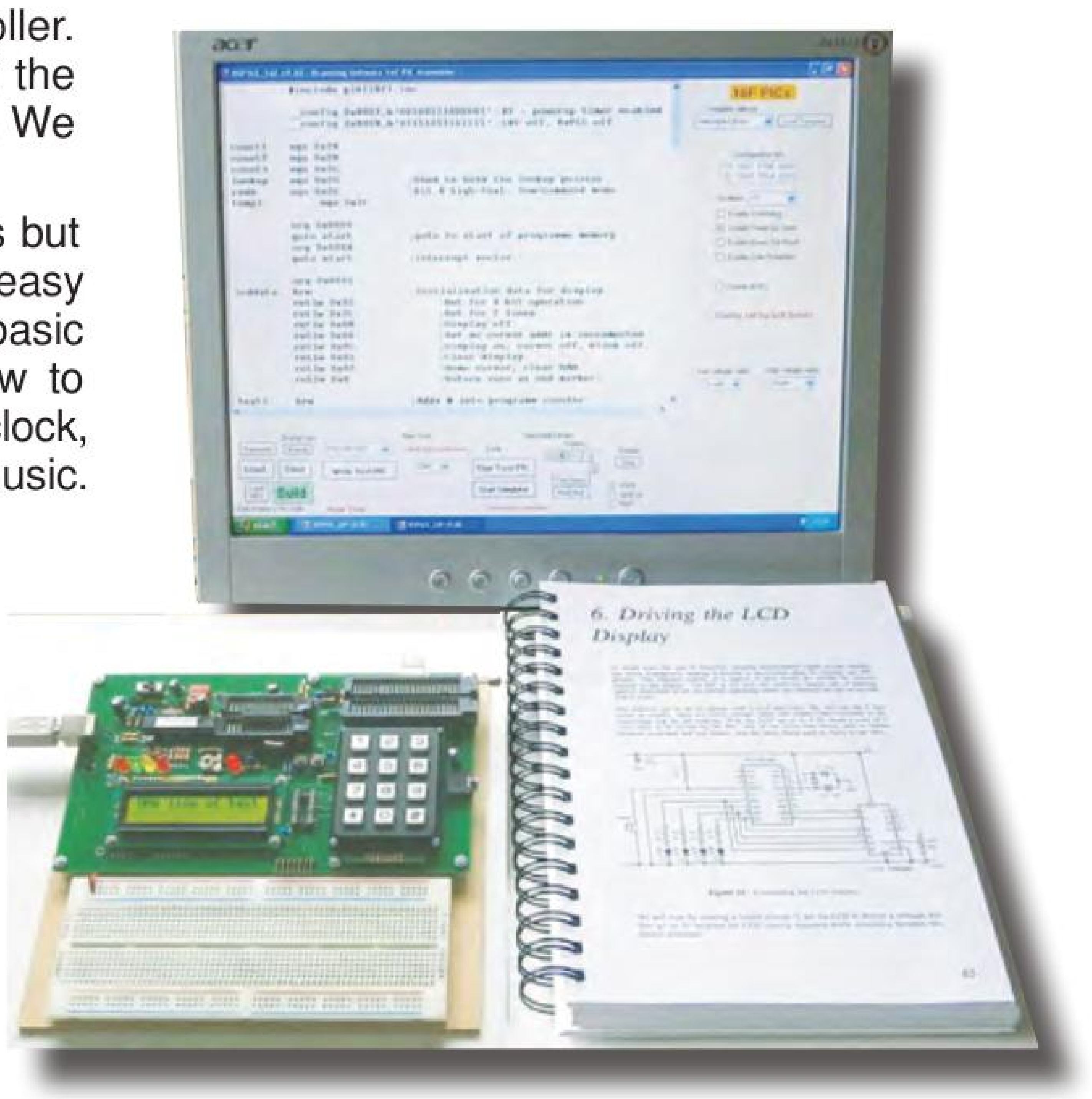


Fig.40. Ceramic cartridge – the two bi-morph elements are visible

The Brunning Software P931 PIC Training Course



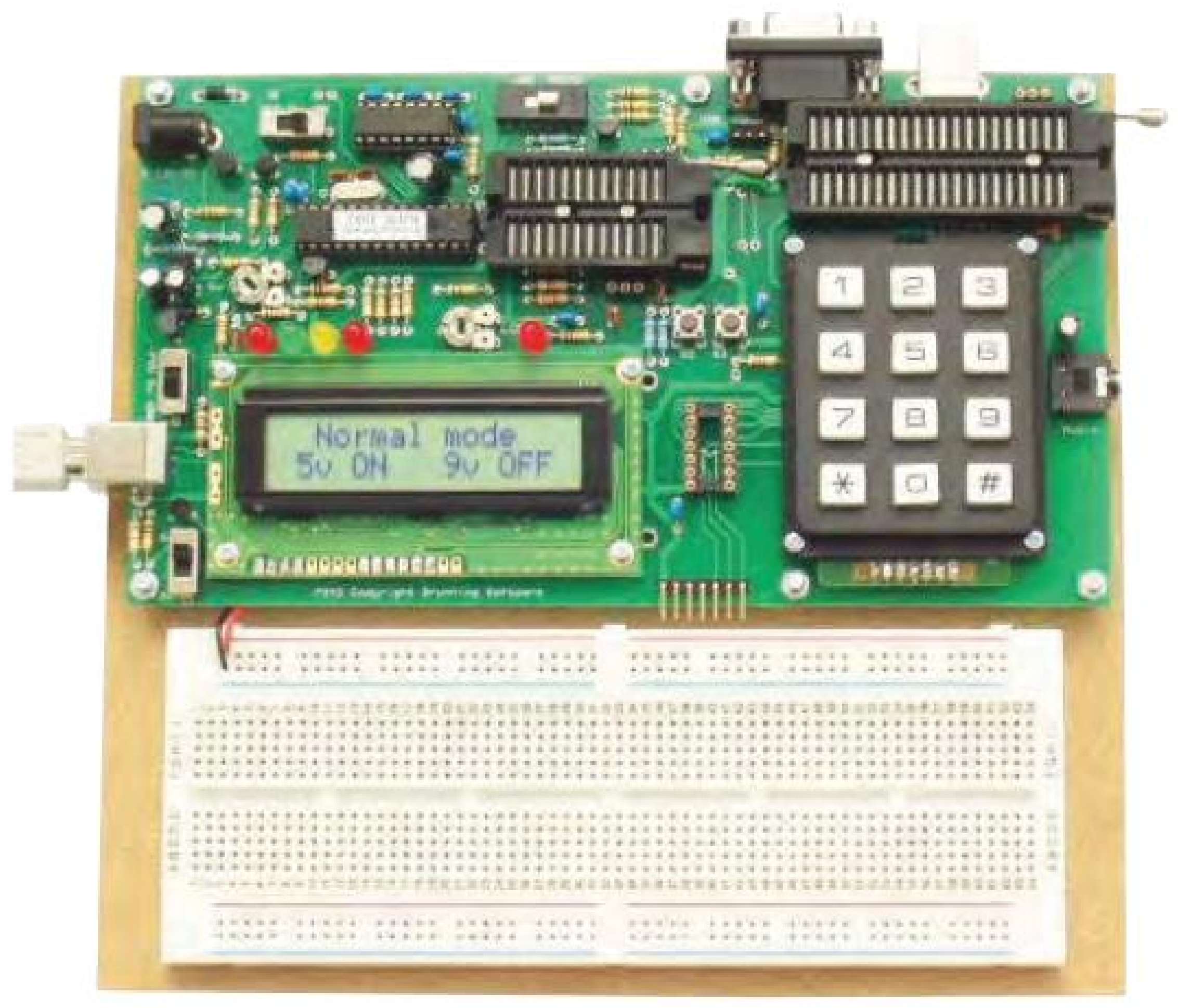
We start by learning to use a relatively simple 8 bit PIC microcontroller. We make our connections directly to the input and output pins of the chip and we have full control of the internal facilities of the chip. We work at the grass roots level.

The first book starts by assuming you know nothing about PICs but instead of wading into the theory we jump straight in with four easy experiments. Then having gained some experience we study the basic principles of PIC programming., learn about the 8 bit timer, how to drive the alphanumeric liquid crystal display, create a real time clock, experiment with the watchdog timer, sleep mode, beeps and music. Then there are two projects to work through. In the space of 24 experiments two project and 56 exercises we work through from absolute beginner to experienced engineer level using the latest 16F and 18F PICs.

The second book introduces the C programming language in very simple terms. The third book Experimenting with Serial Communications teaches Visual C# programming for the PC (not PIC) so that we can create PC programmes to control PIC circuits.

In the fourth book we learn to programme 32 bit MX PICs using fundamental C instructions. Most of the code is the same as already used with the 8 bit PICs so the same experiments are easily adapted. Then life gets more complex as we delve into serial communications with the final task being to create an audio oscilloscope with advanced triggering and adjustable scan rate. (To programme 32 bit PICs you will need to plug on a PICkit3 which you need to buy from Microchip, Farnell or RS for £38).

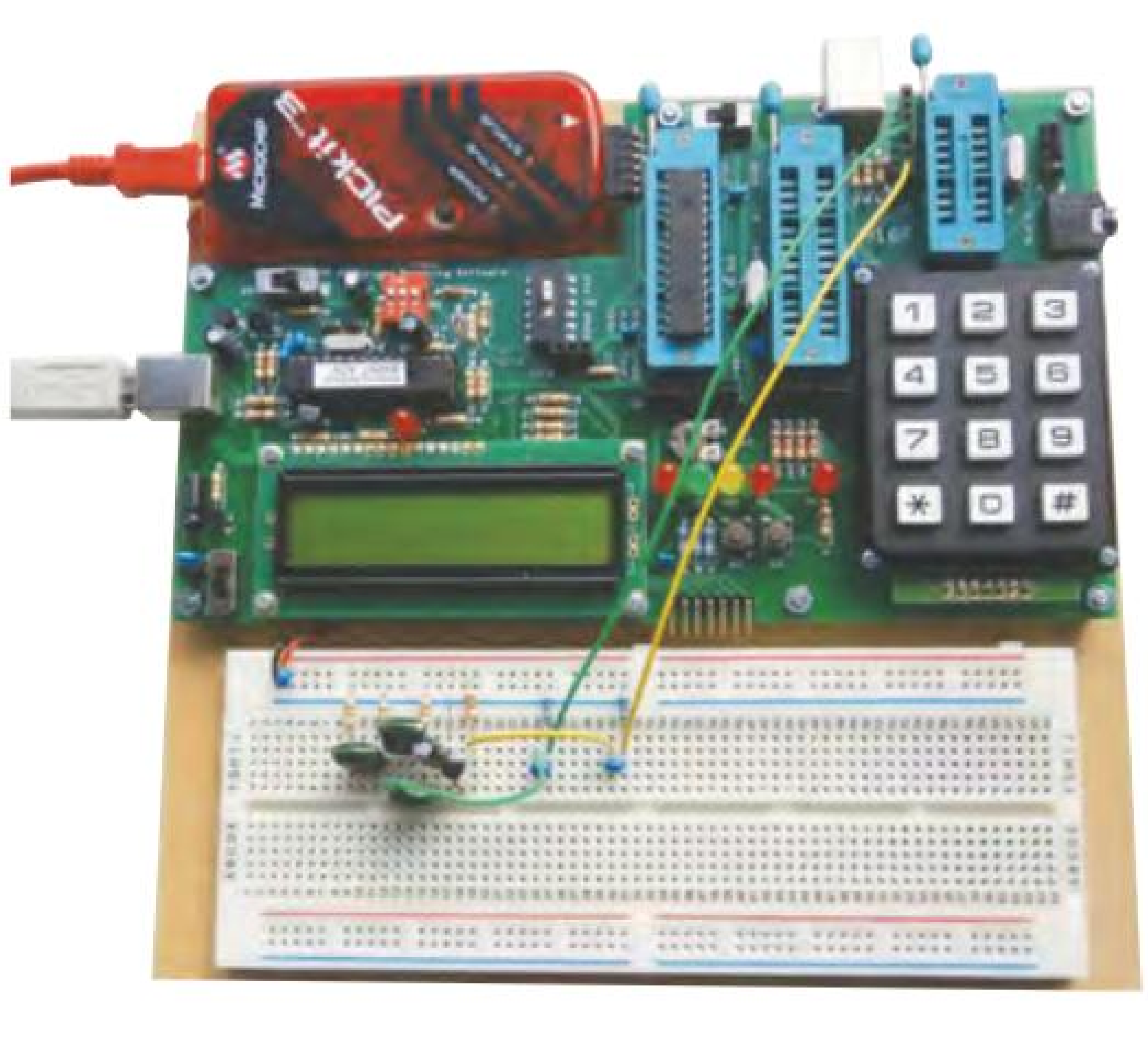
P931 course including UK carriage..... £158.00
Experimenting with Serial Coms..... £31.00
Experimenting with 32 bit PIC C..... £46.00



P942 PIC Training Course

The P942 PIC training course offers exactly the same training as the P931 but the P942 has the ability to verify the data in the PIC at high voltage, normal voltage and low voltage. Verifying that the code in the PIC can be read at high and low voltage is usually recommended if the PIC is to be used in a permanent circuit. For training and experimental work high and low verification is not necessary.

P942 course including UK carriage..... £183.00
Experimenting with Serial Coms..... £31.00
Experimenting with 32 bit PIC C..... £46.00



P955 PIC Training Course

The P955 PIC training course also offers exactly the same training as the P931 but the P955 has a permanent socket for 28 pin 32 bit PICs which makes it more convenient if experimenting with 32 bit PICs is your main interest. (The fourth book Experimenting with 32 bit PIC C can be used with the P931 and P942 using a special adaptor which fits into the 40 pin ZIF socket).

P955 course including UK carriage..... £170.00
Experimenting with Serial Coms..... £31.00
Experimenting with 32 bit PIC C..... £31.00

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Max's Cool Beans

By Max The Magnificent

Tri-colour LEDs – Part 3

As I always say: 'Show me a flashing LED, and I'll show you a man drooling.' This phenomenon is even more pronounced in the case of tri-color LEDs. In my first column on these little scamps (*EPE*, July 2015), I considered a traditional tri-colored device with a common-cathode terminal and three anode terminals – one each for the red, green, and blue sub-LEDs. As part of this, we noted that if we limited ourselves to simply turning these LEDs on and off, then we can achieve 2^3 = 8 different colour combinations (assuming the all-off = black combination is a colour). These colours are black, red, green, blue, yellow (red + green), cyan (green + blue), magenta (red + blue), and white (red + green + blue). The big problem with these little rascals is that *each* tri-color LED requires *three* microcontroller (MCU) pins to drive it.

In Part 2 of this mini-series (*EPE*, August 2015), we discussed driving each LED terminal using a pulse-width modulated (PWM) MCU output. In the case of the Arduino, we have 8-bit PWMs, which theoretically allows us to achieve 2^8 = 256 different brightness values, from 0 (fully off) to 255 (fully on). Since we can do this for each sub-LED, this means that we can theoretically achieve 2^8 * 2^8 * 2^8 = 16,777,216 different colours. This is all jolly nice, but we are still faced with the fact that each tri-color LED requires three micro-controller (MCU) pins to drive it, which does tend to limit you as to what you can achieve. Happily, there is a way by which we can drive hundreds of tri-coloured LEDs using a single MCU pin... so, how do we do that? – read on!

Introducing the WS2812

A couple of years ago, a company called Worldsemi introduced a device called the 'WS2812'. This little beauty is presented in a 5mm × 5mm square package that contains super-bright red, green, and blue LEDs, along with an associated control chip and current-limiting circuitry.

The great thing about these devices is that they have only four pins: 0V (ground), 5V (power), Data-In, and Data-Out. This means that they can be daisy-chained together, with the MCU driving the Data-In of the first element, the Data-Out from the first element driving the Data-In of the second, and so forth, as illustrated in Fig.2.

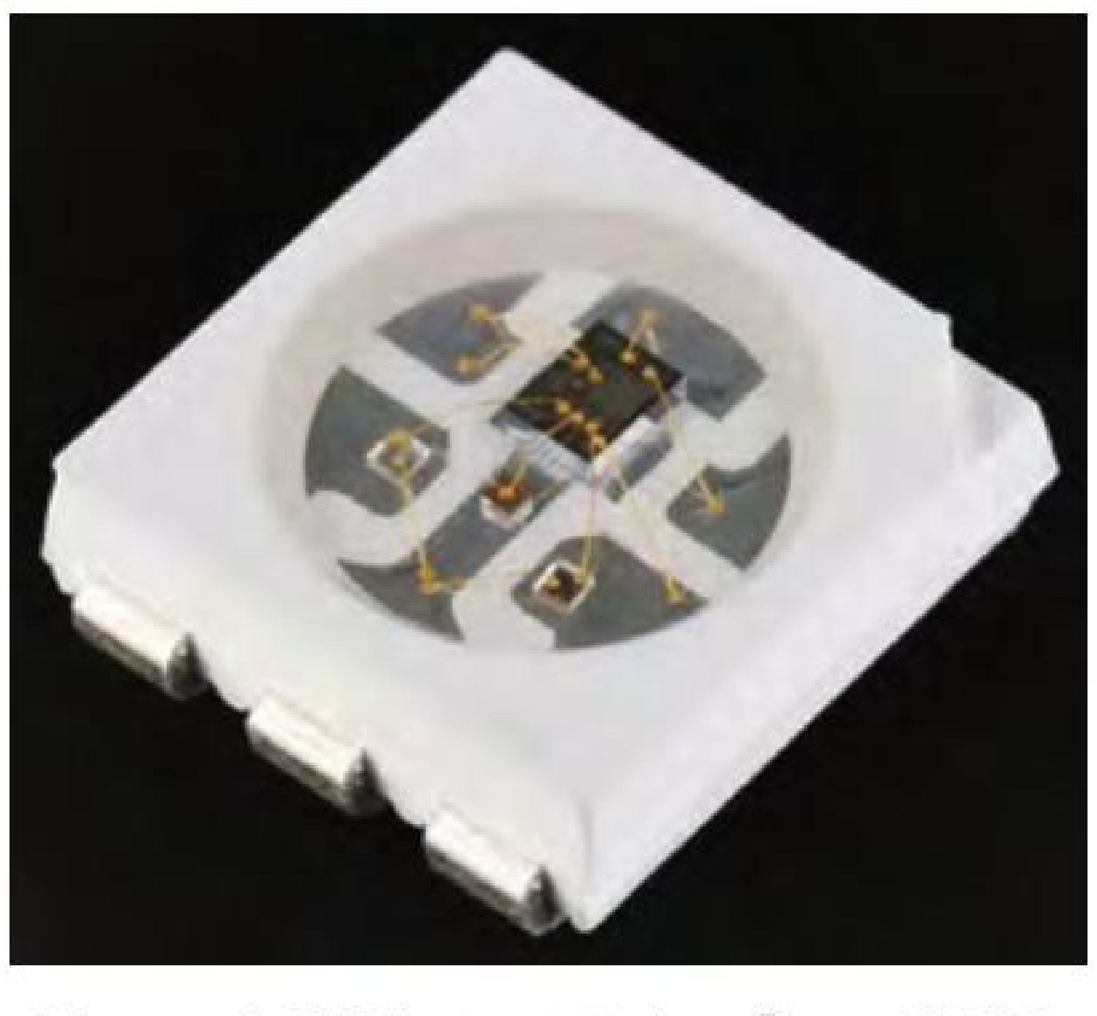


Fig.1. A WS2812 tri-colour LED

Different companies have used these devices as the basis for a wide variety of products. For example, the folks at Adafruit have a whole range of WS2812-based products that they call NeoPixels. These range from strips to rings to panels to ... all sorts of things – see: **http://bit.ly/1gr71a3**.

I've been using NeoPixels a lot in my hobby projects, such as my *Bodacious Acoustic Diagnostic Astoundingly Superior Spectromatic* (*BADASS*) display and my *Vetinari Clock* – check out these YouTube videos, respectively at: **http://bit.ly/1I5ncTc** and **http://bit.ly/195JKWR**. The great thing about Adafruit is that they give you a NeoPixel software library along with a bunch of examples on how to use it.

One thing that can confuse beginners is that the Adafruit library includes a function called `setPixelColor()`, but this doesn't actually update the physical pixel itself. Take my *BADASS* display, for example. Suppose we use the Adafruit library to declare a string of 256 NeoPixels called `myNeos` (these will be numbered from 0 to 255). Now suppose I use the following statement, where the three parameters of 255 are associated with that pixel's red, green, and blue channels, respectively:

```
for (int i = 0, i < 256; i++)
  myNeos.setPixelColor(i,255,255,255);
```

What do you think will happen on the display? The answer is 'not a lot.' This is due to the fact that, when we first declare our NeoPixels, the system creates an array in the Arduino's memory. This array contains one main element for each NeoPixel, and this main element is comprised of three byte-sized sub-elements to contain that pixel's red, green, and blue values. The `setPixelColor()` function actually modifies the values in this array. When we are ready to update the display, we would use the `show()` function; for example:

```
for (int i = 0, i < 256; i++)
  myNeos.setPixelColor(i,255,255,255);

myNeos.show();
```

The `show()` function streams the entire contents of the array from the Arduino's memory into the NeoPixel string. In the above example, this means that all of the pixels will turn fully on simultaneously. Now, assuming that all of the pixels are already on, how would we turn them off one at a time with a tenth of a second (100ms) delay between each one? Well, we could achieve this as follows:

```
for (int i = 0, i < 256; i++)  {
  myNeos.setPixelColor(i,0,0,0);

  myNeos.show();
  delay(100);
}
```

As you can see, we've moved the `show()` function inside the `for()` loop, and we've also added a `delay()` function inside the loop so that we can see the pixels being extinguished one after another.

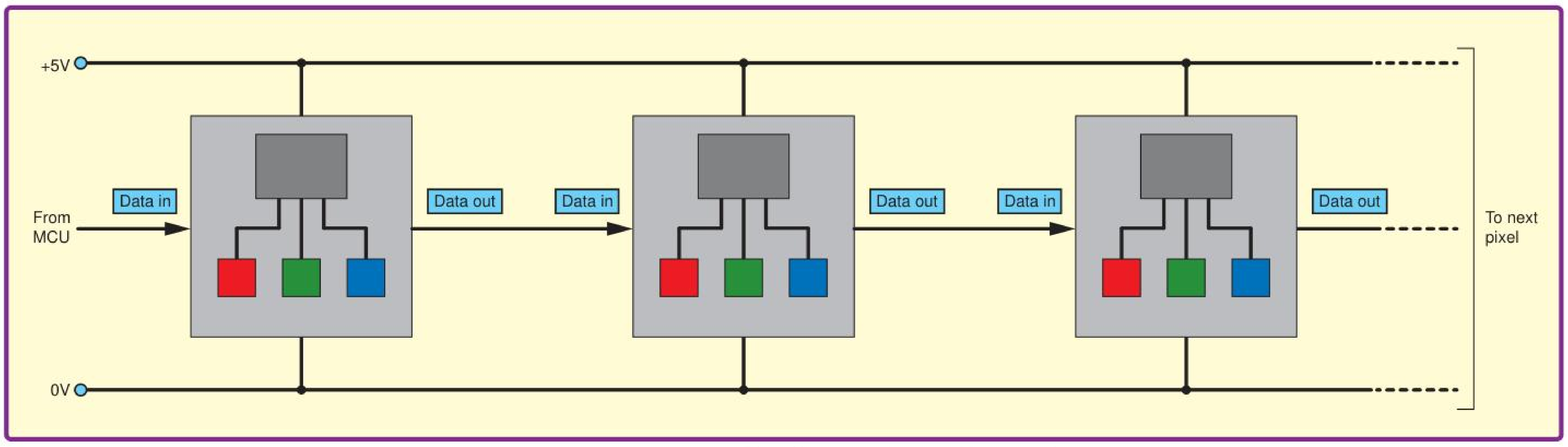
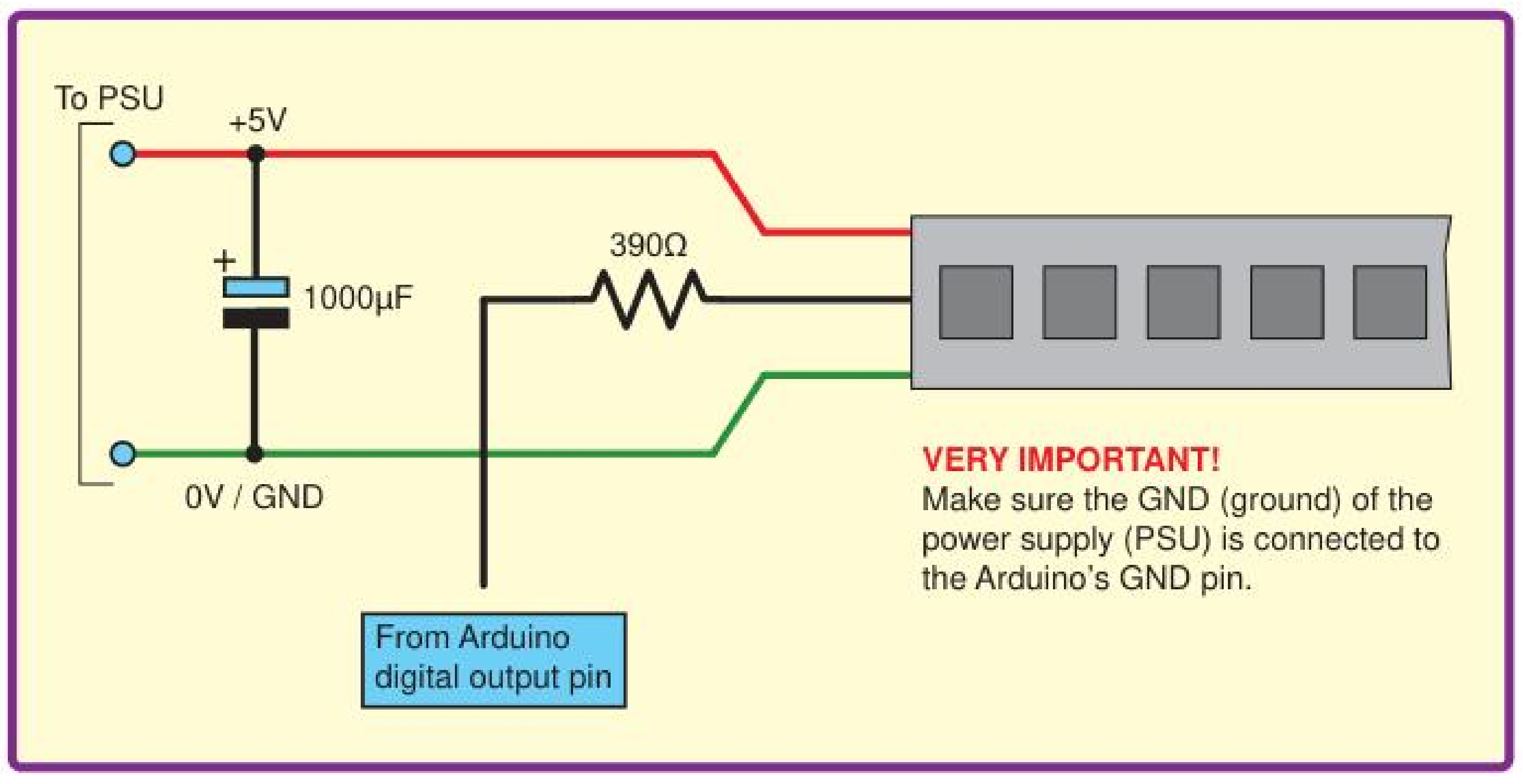


Fig.2. WS2812s can be daisy-chained together



VERY IMPORTANT!
Make sure the GND (ground) of the power supply (PSU) is connected to the Arduino's GND pin.

Fig.3. Using a separate supply for the NeoPixels

How much power?!

Now, one thing you have to consider is how you're going to power the little ragamuffins. Your knee-jerk reaction may be along the lines of 'LEDs are very efficient and they don't consume much power.' Well, that's as may be, but the power starts mounting up when you use a lot of LEDs. Adafruit's website states that each NeoPixel can consume up to 60mA when its red, green, and blue channels are all fully on. In the case of my *BADASS* display, I have a 16 × 16 = 256 array of NeoPixels, which equates to 256 × 60mA = (approx) 15.5A (I actually use a 20A supply to give myself some 'wriggle room').

You may decide to power your NeoPixels using a separate power supply to that used to power your Arduino. In this case, it's very important to ensure that the ground (GND, 0V) of this power supply is connected to the Arduino's GND pin. It's also a good idea to mount a 1,000µF electrolytic capacitor across the power supply rails, as illustrated in Fig.3.

Furthermore, it's a good idea to insert a resistor in series with the data signal coming from the Arduino. This resistor, which is used to dampen overshoot and undershoot on the signal, should be mounted close to the first element in the NeoPixel chain (ie, *not* at the Arduino end of the signal path). Adafruit's website says the value of this resistor should be 300Ω to 500Ω; I typically use a 390Ω resistor. (NeoPixels are also prone to being damaged by electrostatic discharge (aka ESD) so you should take appropriate earthing precautions while using them.)

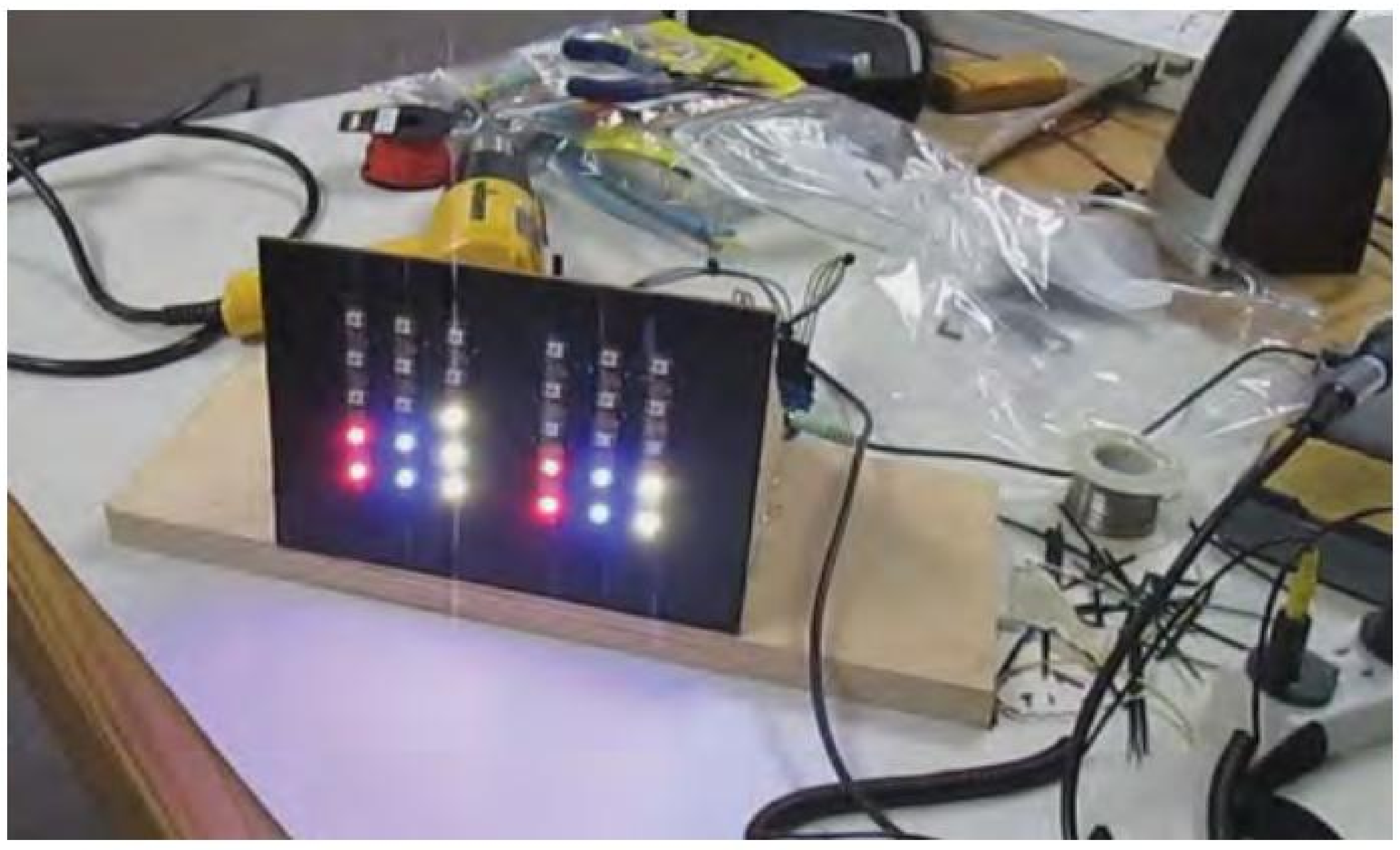


Fig.4. Jacob's mini-BADASS *display*

In the case of my projects, I typically modify my Arduino so that it works from a 5V supply (the unmodified Arduino requires a 7V to 12V supply, which is subsequently regulated down to 5V), and then I use the same regulated supply to power both the Arduino and the NeoPixels. However, this involves removing and bypassing the Arduino's on-board regulator and tweaking a few resistor values, so it's not for the faint of heart.

Using the Arduino 5V pin

The other alternative is to power the NeoPixels from the Arduino's own 5V supply pin, but this means you are limited as to how many NeoPixels you can use. For example, I'm teaching a young lad called Jacob about using microcontrollers – he comes into my office for an hour each week. Not surprisingly, Jacob would like his own *BADASS* display, so we built him a mini version, as shown in Fig.4 (check out this YouTube video **http://bit.ly/1UalHbP**).

Now, Jacob powers his Arduino from his PC's USB port, which is only specified to provide 0.5A (500mA). When running at full pelt, an Arduino Uno R3 draws around 45mA, and the spectrum analyzer shield only consumes about 3mA, so if we assume a worst-case of 50mA, this leaves us with 450mA to play with.

Although Adafruit says that each NeoPixel can draw up to 60mA when fully on (20mA each for the red, green, and blue sub-pixel LEDs), we've actually put a string of NeoPixels on the test bench and measured them as drawing an average of 43mA full on, which we rounded up to 45mA (or 15mA per color). So... in the case of Jacob's mini display, we calculated that if we have three columns per channel – one each for the base, middle, and treble – and if each column has five NeoPixels – and if each NeoPixel is used to display only one colour (red, green, or blue), then we end up with 6 × 5 × 15mA = 450mA (I love it when a plan comes together).

'Ah ha!' I can hear you say, 'but Fig.4 shows you using red, blue, and yellow (which is composed of red and green) – how can this be?' The answer is really simple – so simple, in fact, that I'll leave you to work it out for yourselves!

Any comments or questions? – please feel free to send me an email at: **max@CliveMaxfield.com**

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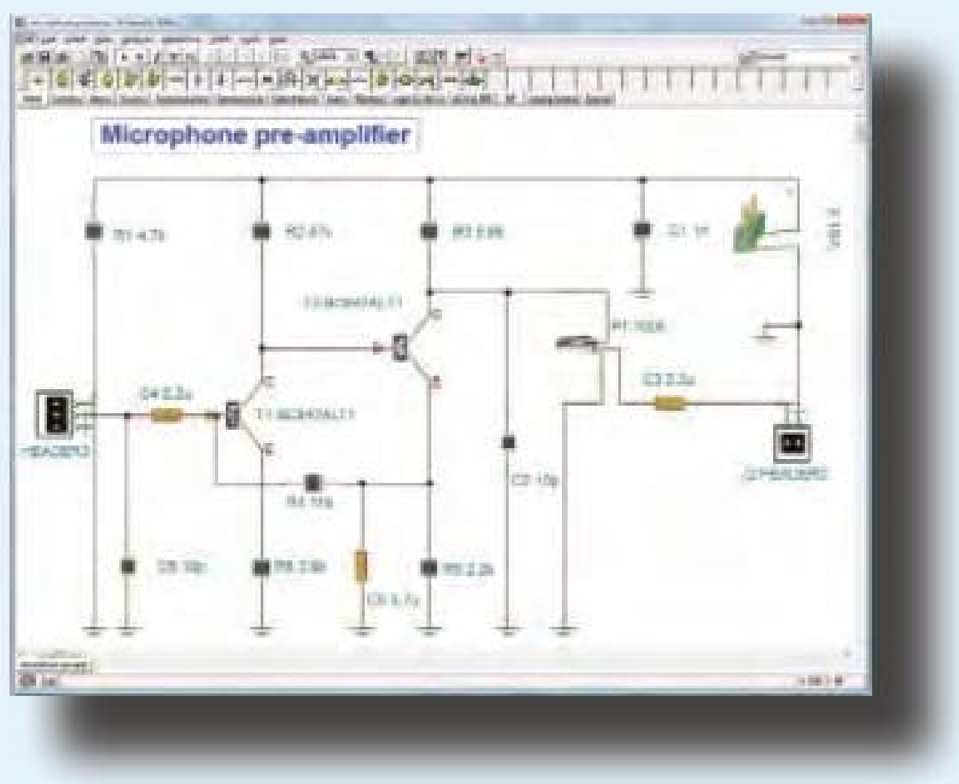

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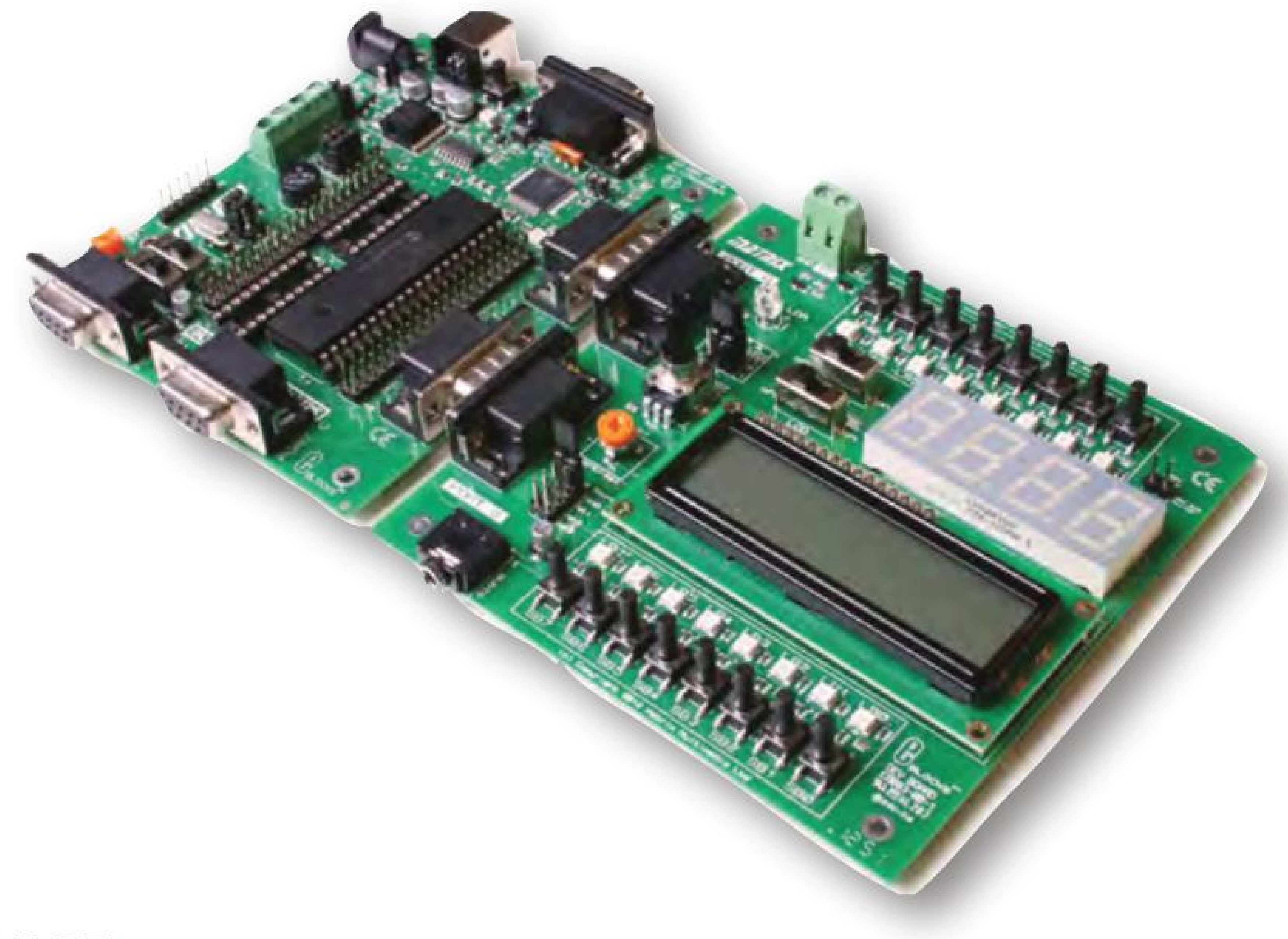

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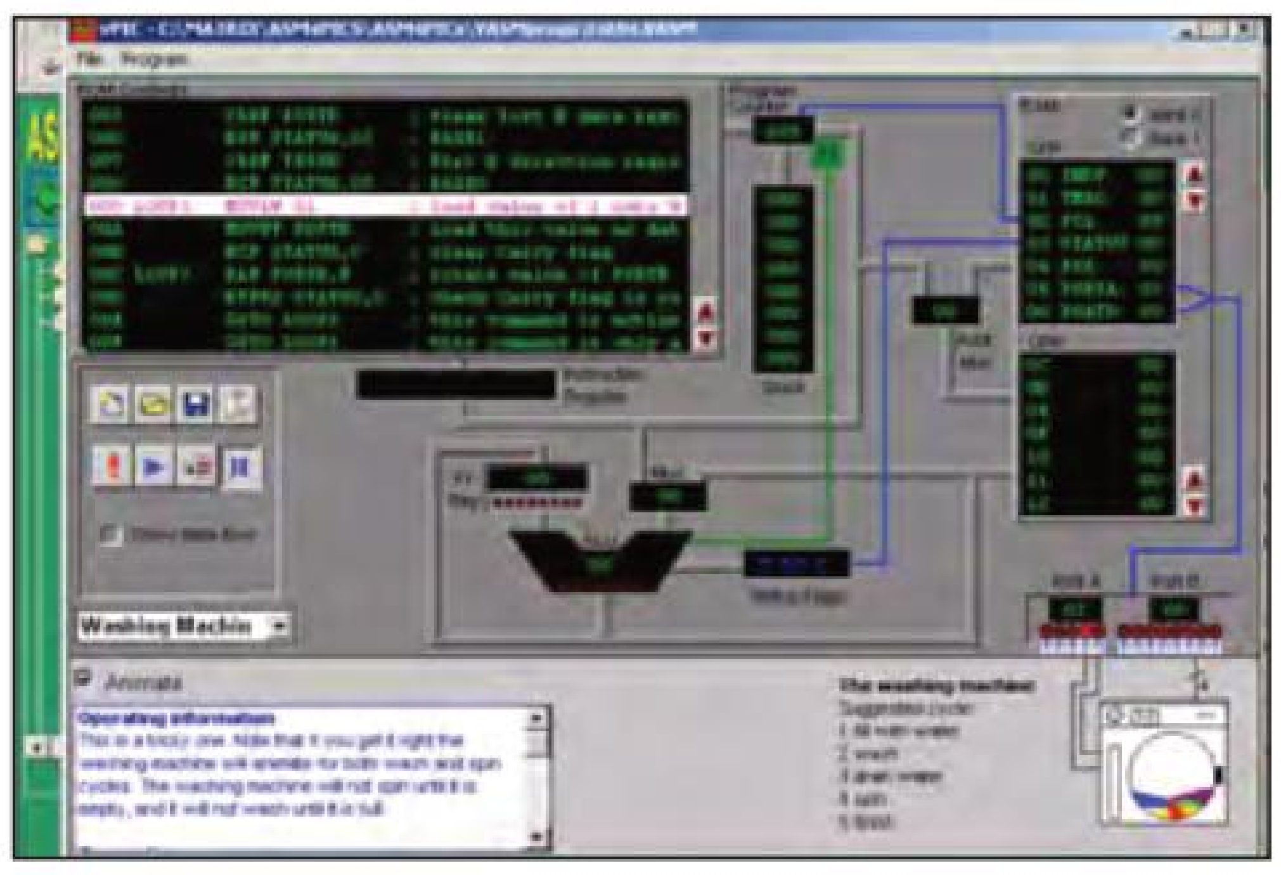

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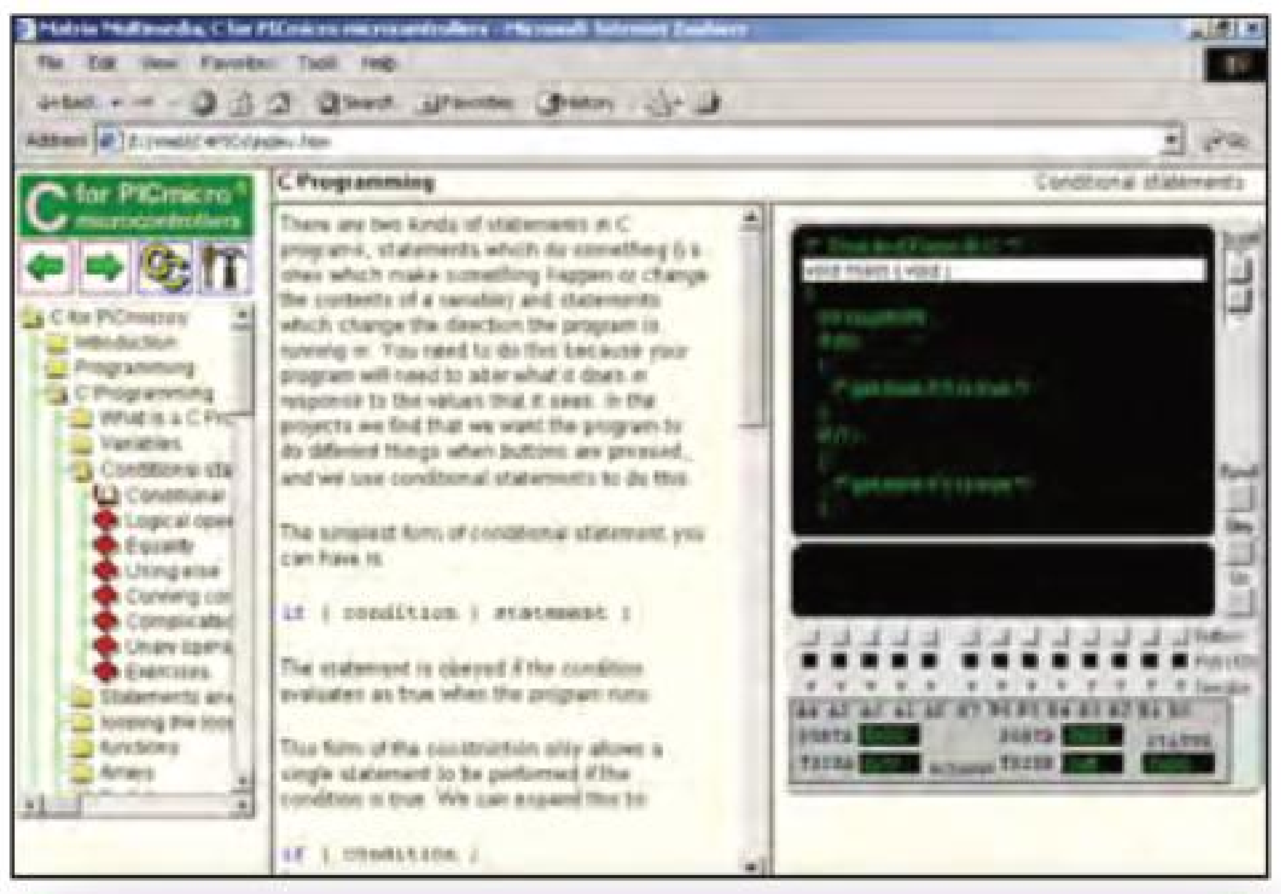

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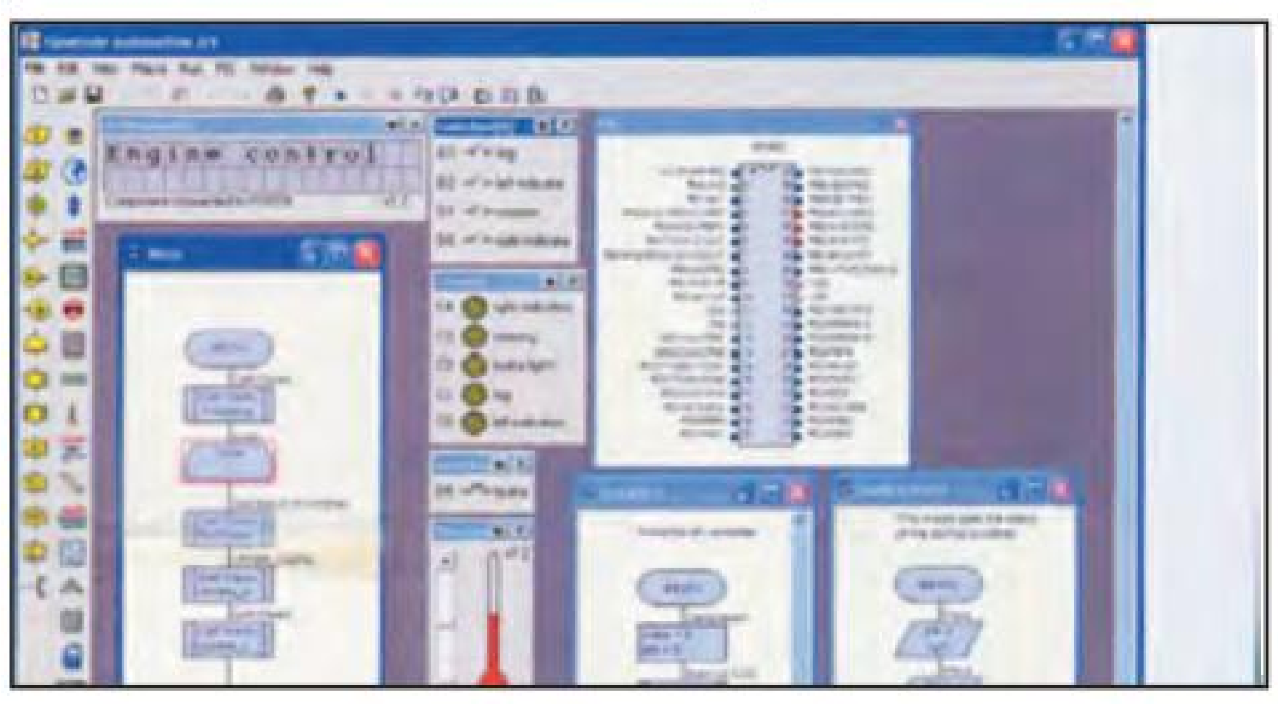

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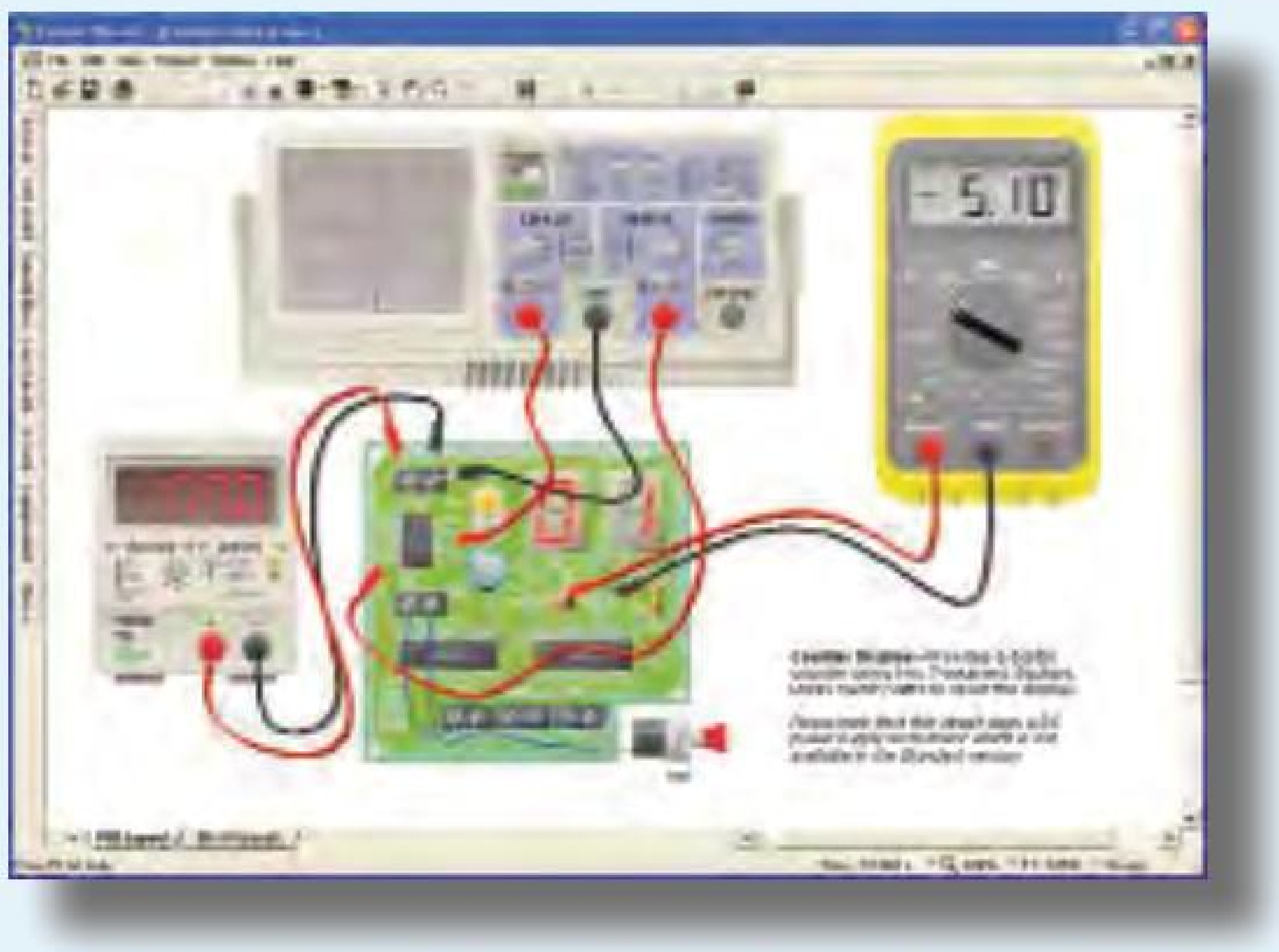

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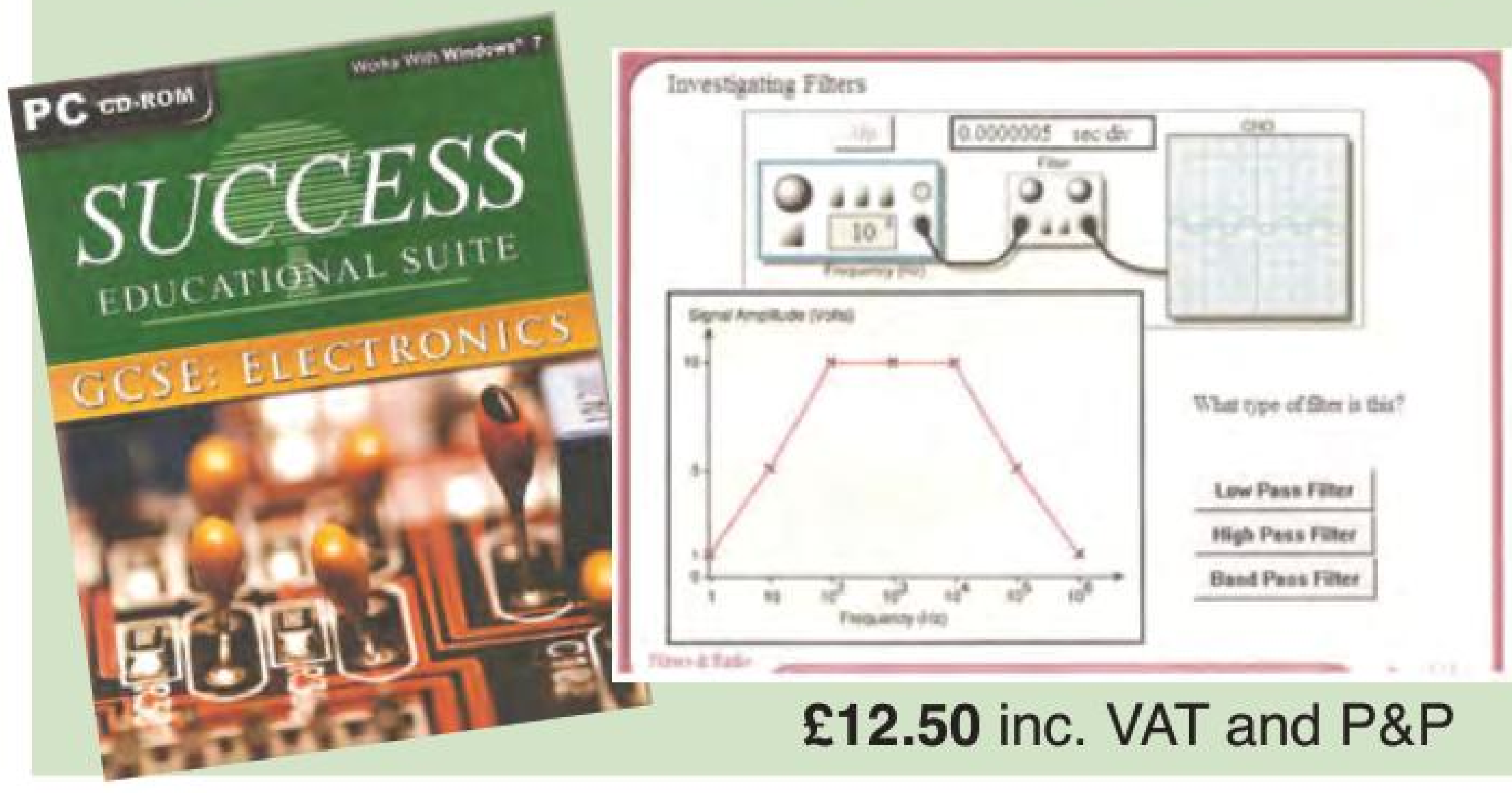

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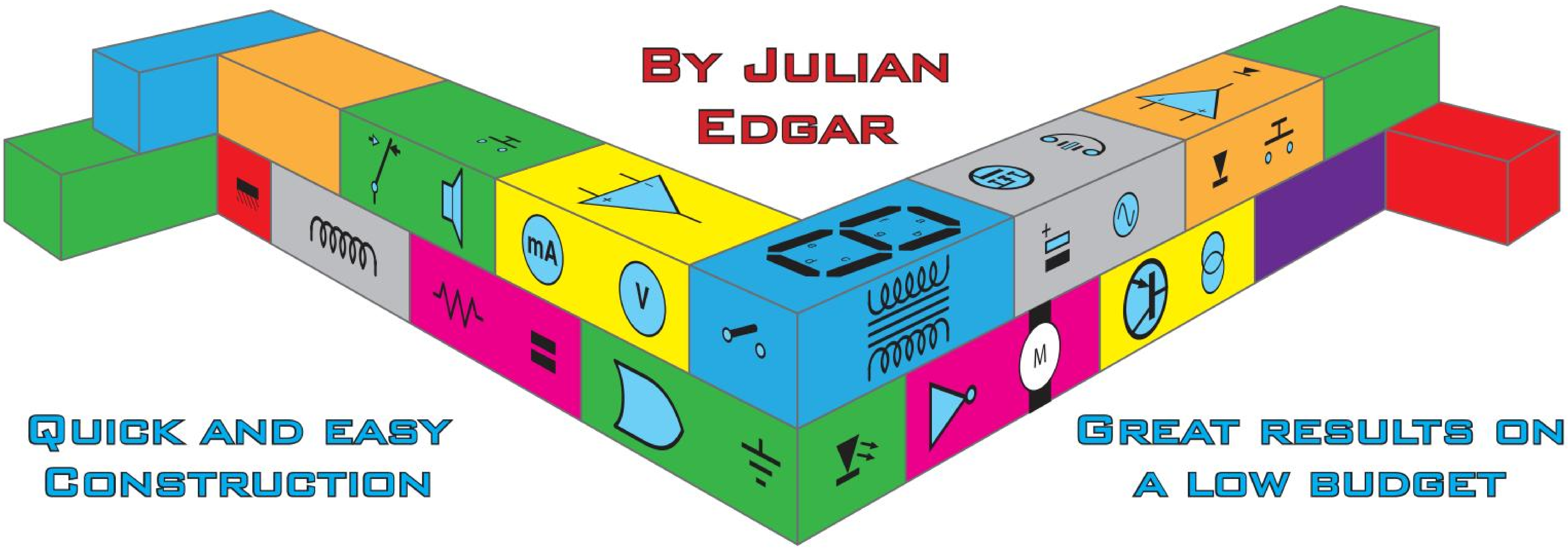
ELECTRONIC BUILDING BLOCKS



BY JULIAN EDGAR

QUICK AND EASY CONSTRUCTION

GREAT RESULTS ON A LOW BUDGET

9W STEREO AMPLIFIER

Large complex projects are fun, but they take time and can be expensive. Sometimes you just want a quick result at low cost. That's where this series of *Electronic Building Blocks* fits in. We use 'cheap as chips' components bought online to get you where you want to be... FAST! These projects range from around £15 to under a fiver... bargains!

9W Stereo Amplifier

Need a cheap audio amplifier – perhaps as a test bench amplifier or for a kid's MP3 sound system? Well it doesn't get much cheaper than this – try under £4 for the prebuilt amplifier module, including postage to your UK letterbox!

So what do you get for your measly £4? Well, it isn't exactly 'Hi-Fi' – but then again, neither is it anywhere near as nasty as you might first think! (See Fig.1) The module uses a Tripath TA2024 integrated circuit that delivers 9W per channel into a 4Ω load at a total harmonic distortion of 0.04%. Crank it up to 15W per channel and distortion rises very substantially to 10% – so you don't want to overdo it! Dynamic range is 89dB. The module comes fully built and is 92mm x 53mm in size.

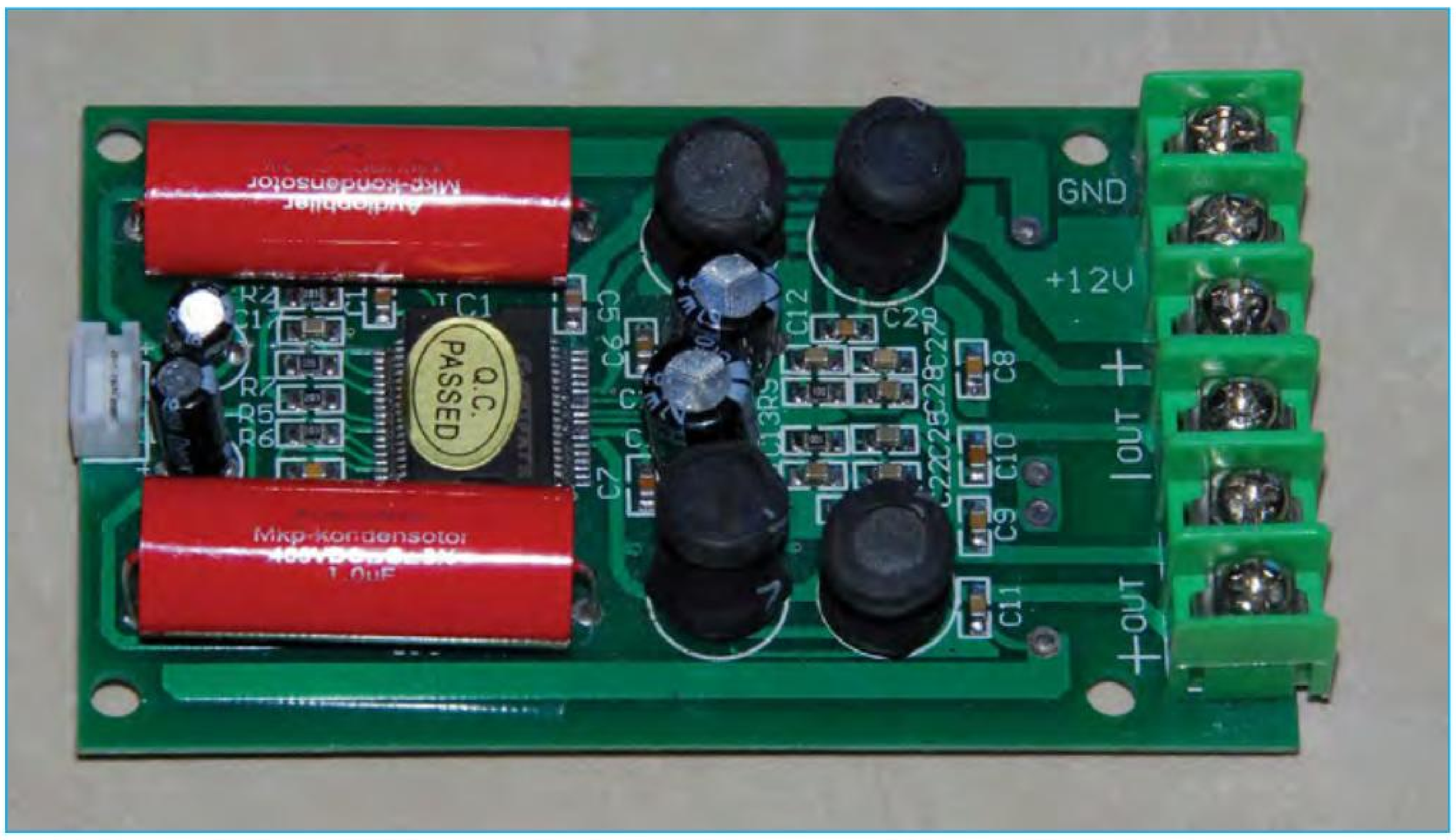


Fig.1. Fantastic value – a 9W stereo amplifier for under £4. It won't challenge high-end Hi-Fi, but performs respectably when driven adequately and paired with a couple of 4Ω speakers

Connections

At one end of the module is a terminal strip to allow connection of the two speakers, and power and ground (Fig.2). The board is spec'd at 12V, but it will work down to 7V. Note that it is *not* reverse-polarity protected.

At the other end of the board is a three-terminal connector for left, right and ground audio inputs (Fig.3). A short input cable is supplied with an appropriate plug and lead.

At 10W audio output power, the manufacturer rates the efficiency of the IC at about 78% (with 4Ω speakers), so the chip does not need a heat sink. In fact, in our testing, it stayed relatively cool even when running continuously at full power on audio material. The cheapest and easiest approach to powering the module is to use a discarded mains-powered plugpack. Any design with a DC output of around 8-12V will work fine.

Driving the amp

Many people will want to run the module from a phone or other MP3 player. Cheapest is to get an old pair of earphones (or any other discarded cord) that has a 1/8-inch stereo plug already wired into place. Just cut off the lead and make the connections to the input cable supplied with the module (Fig.5).

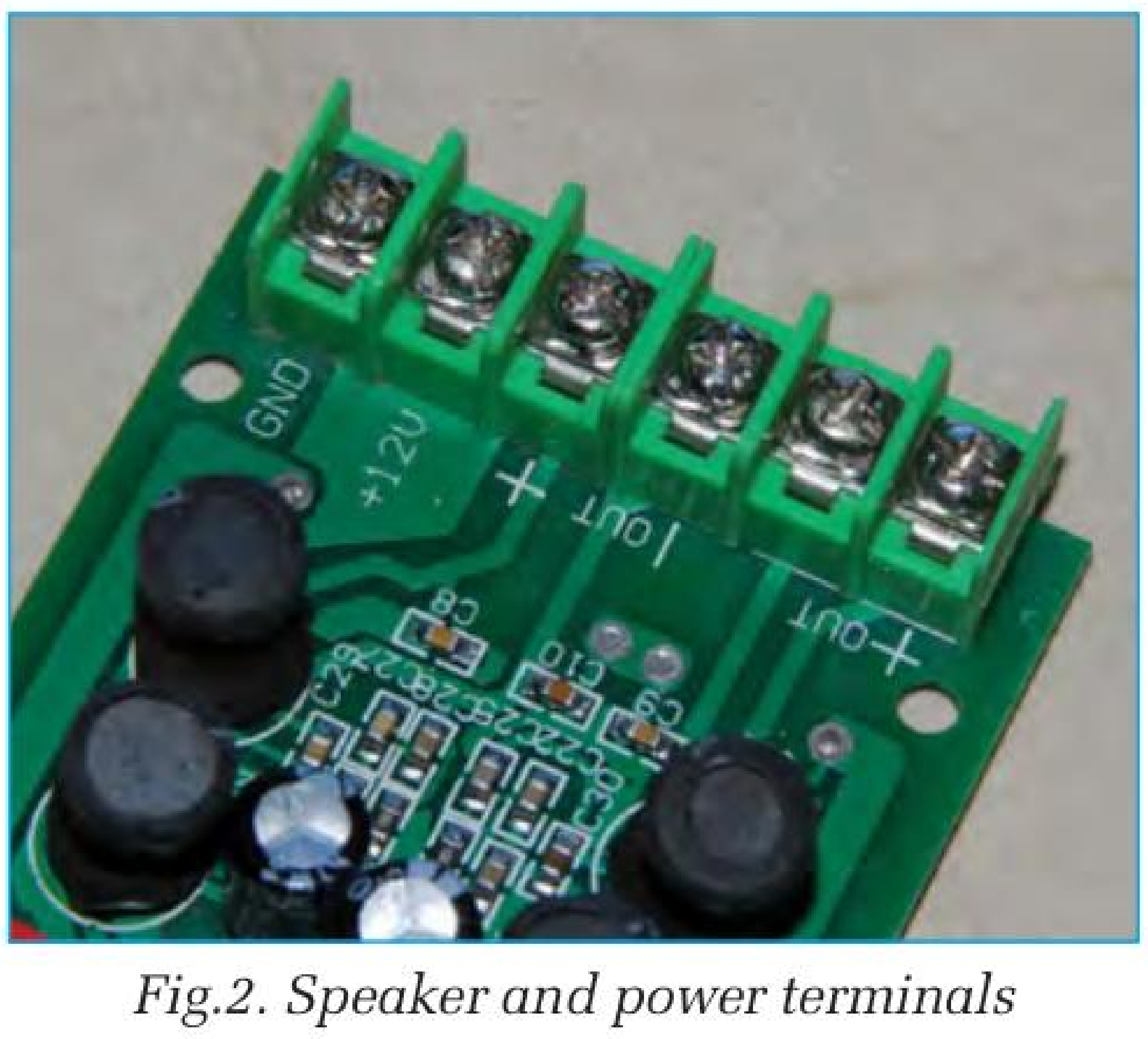


Fig.2. Speaker and power terminals

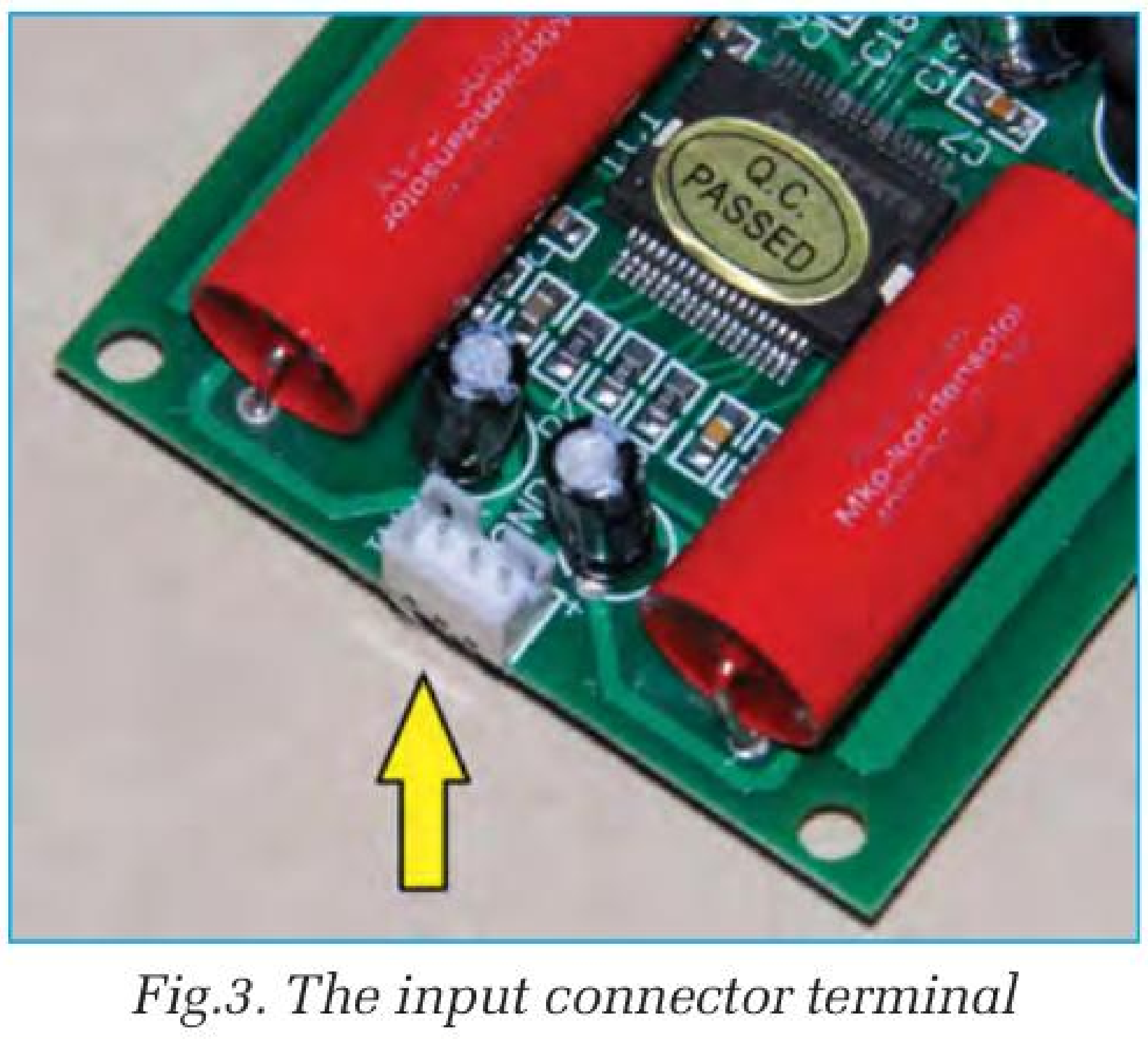


Fig.3. The input connector terminal

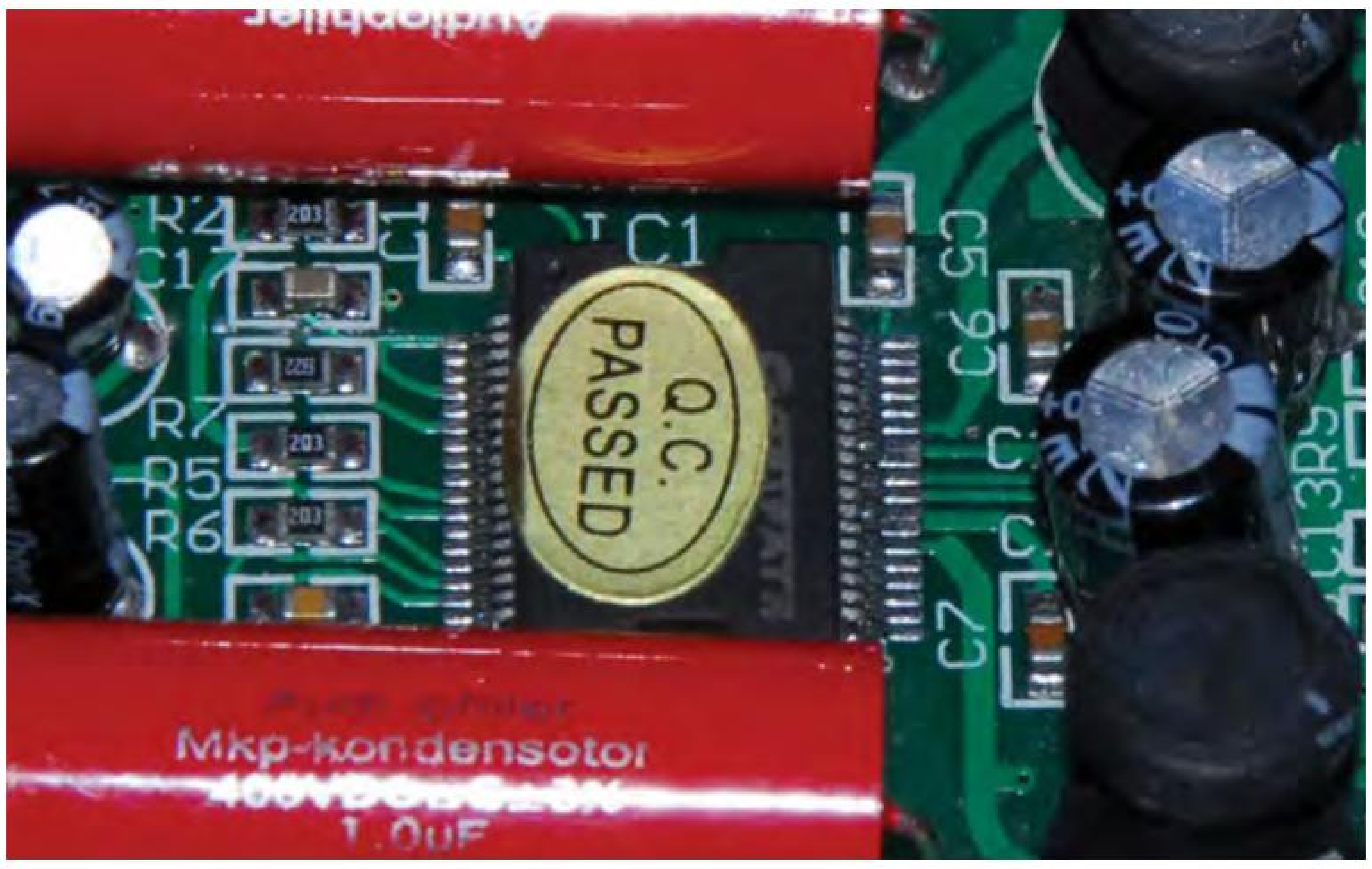


Fig.4. The heart of the beast – a Tripath TA2024 IC

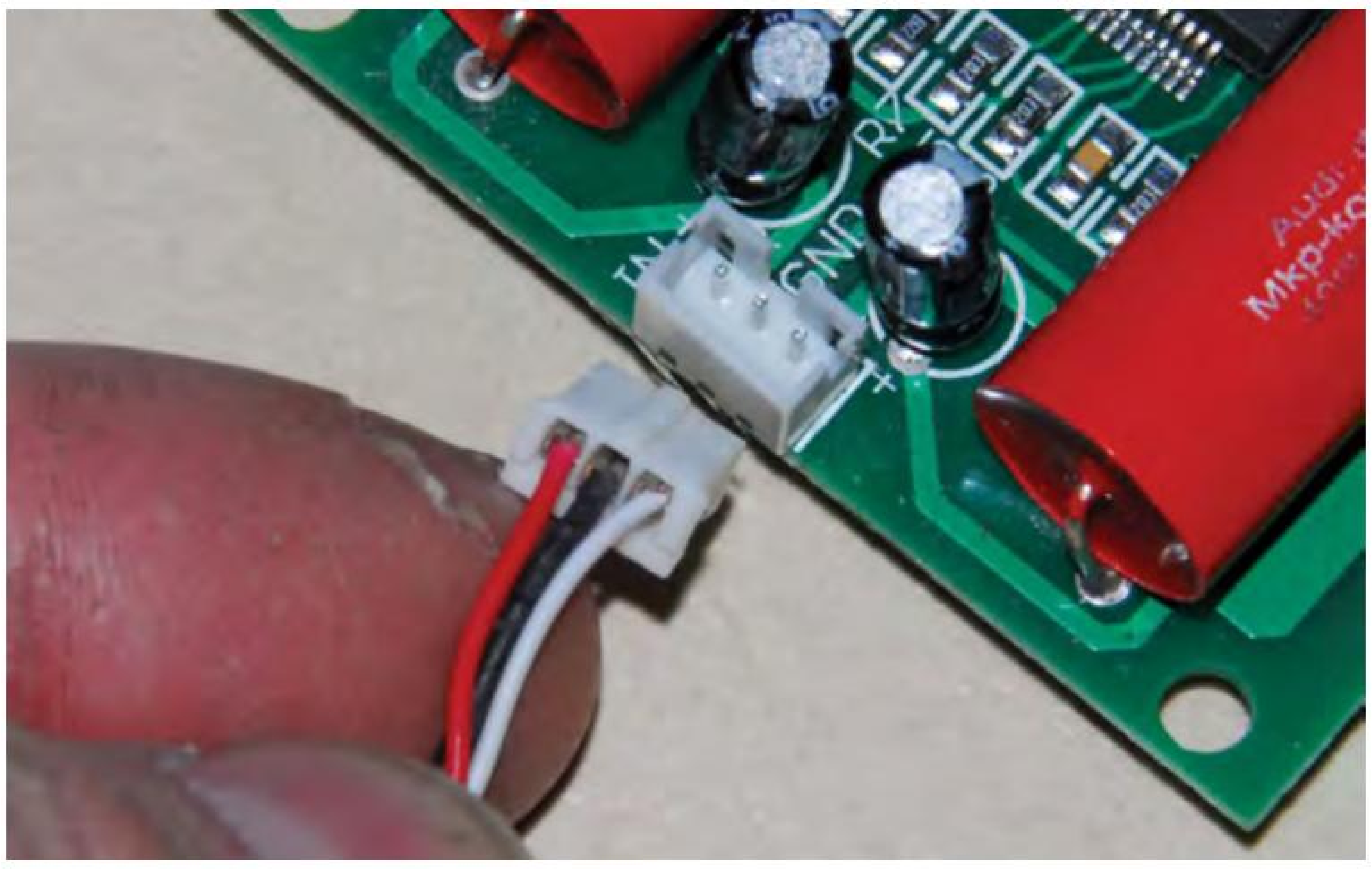


Fig.5. Connecting up the input

The conductor that connects to the tip of the plug goes to the red input wire of the module; the central conductor on the plug to the white input lead; and the base of the plug to the black input. Connect speakers to the appropriate screw terminals, switch on power and listen to music! The output level of an iPhone that we used to test the module worked well – the amp input was not overloaded at even full phone output.

Battery powered?

What about running the module off a 9V battery? At 9V and driving both channels from an iPhone, we measured an average current draw of just 60mA at a low listening level, 80mA at a moderate level, and 185mA at a loud level. That makes battery operation entirely feasible.

Conclusion

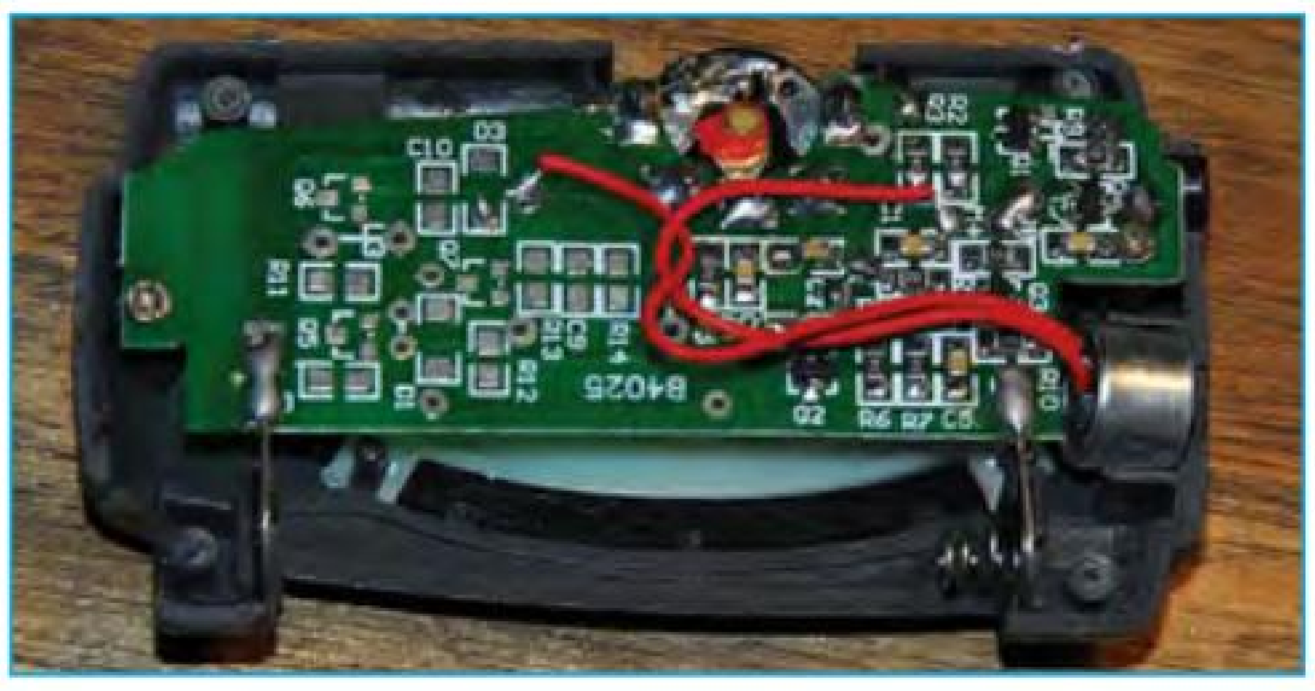
As I said at the beginning, don't even pretend that this is a full-blooded Hi-Fi amplifier that will shake the house. But where you need a utility low-powered amplifier, at just £4 plus an old plugpack, it's pretty hard to go wrong.

Sourcing

At the time of writing, a typical online unit is eBay item 351092217368 at a very reasonable £3.88 incl delivery.

Next month

Fancy an electronic stethoscope? Here is a really neat device that will allow you to listen to all sorts of strange noises in machines and other mechanical devices. Just add some headphones and shielded microphone cable. All this is in our next *Electronic Building Block* article.



Next month – an electronic stethoscope

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READOUT

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All letters quoted here have previously been replied to directly

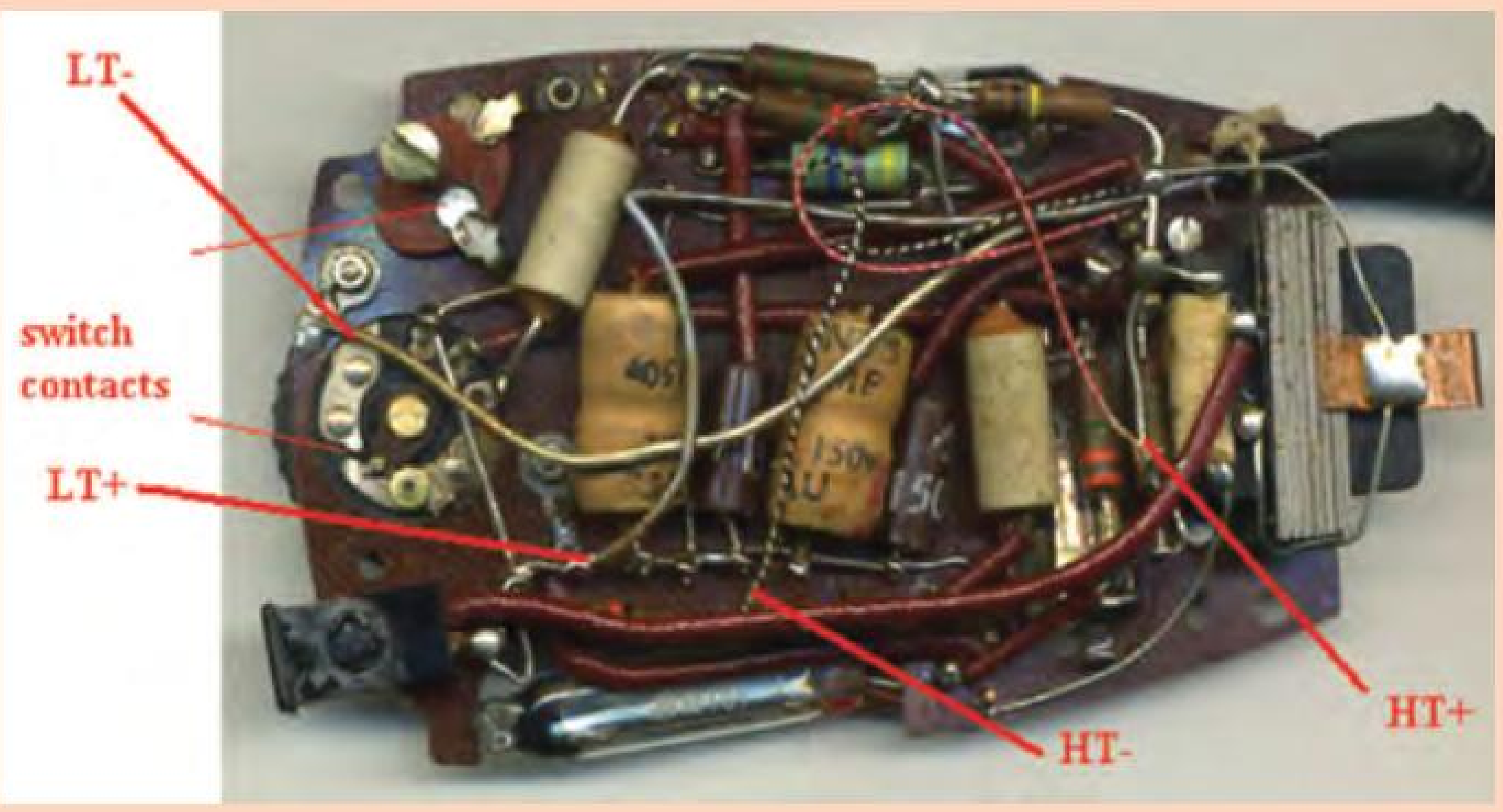
Email: editorial@wimborne.co.uk

☆ LETTER OF THE MONTH ☆

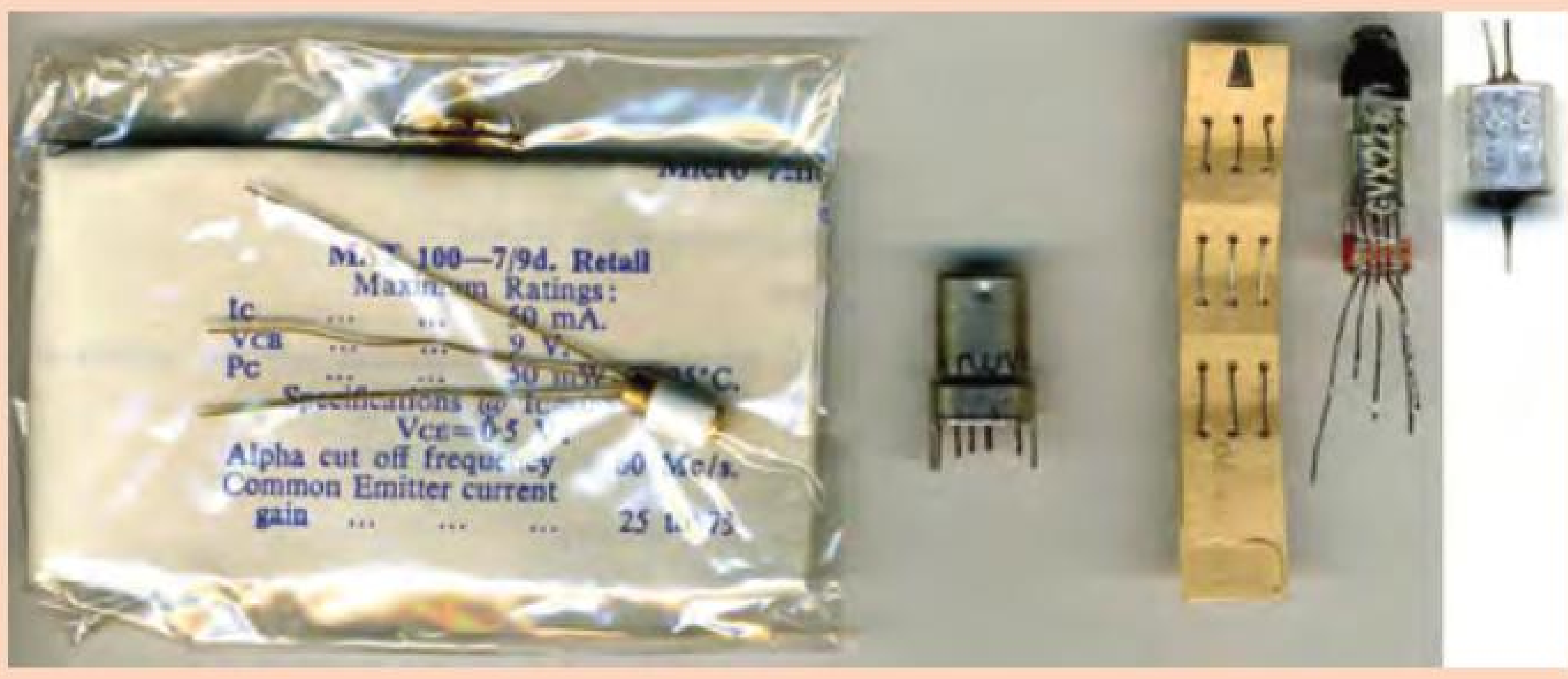
Valve-based hearing aid

Dear editor
There's no need to '...imagine a valve hearing aid!' (*EPE*, June 2015, *Audio Out*, Ge-mania – Part 3, p.52) because here's a picture of me modelling one (my thanks to Christine Mlynek for the photo). The Medresco (MEDical RESearch COuncil) OL35A was one of the first NHS hearing aid patterns. The OL15A is slightly earlier, but the differences are subtle. I haven't an ear moulding so the earpiece is hooked loosely round my left ear. The dangling leads are for the heater and HT ('high tension') batteries, again I have no originals, but they were probably larger than the aid itself. It does work, an AA cell with a touch of series resistance provides the heater supply and a 23A compact 12V battery is sufficient for the HT, even though slightly lower in voltage than specified. Remember that the zinc-carbon cells of the day dropped from their rated potential difference quite quickly, and so circuits had to be tolerant of a lower supply. Also shown is a photo of the circuitry.

The sub-miniature valves are wire-ended and originally intended for guidance/fusing in late wartime ordinance. This is far smaller than the more commonly accepted miniature B7G and B9A valves. The remaining picture shows one of the wire-ended valves (second from right) with an OC59 (still in original package), as also seen in Jake's Ge-mania, Part 3 article in Fig.26 – the transistor replaced valves in smaller aids. Valve designers were aiming to make valves smaller still, one more place to the left is a 6CW4 Nuvistor in its shiny metal can.



Jake mentions Sinclair's transistors rescued from disposal by Plessey. Clive Sinclair claimed to have involvement in the novel microalloy devices – they were actually the poor-spec Plessey throw-outs with a crude paper label bearing his own non-standard MAT type numbers (a MAT100 still in its packet is seen at the far left of the photo). Transistors didn't start this way, of course. The very first offerings were point contact, as exemplified by the rare GET1 seen at the far right of the picture – sadly, non-working (do not confuse with GE numbered devices).



Tumbling audio caps!

On a completely different tack, I've just had to replace the motor-run capacitor in my tumble dryer (which stopped tumbling when the capacitor failed for no obvious reason!). What's that to do with audio? These are non-polarised (not electrolytic) caps in a range of values in the several μF class. ESR measurements are very low. Lateral thinking: could they be used for audio? They're big, but not massive. Available in a metal screening can which is isolated from the terminals.

Godfrey Manning G4GLM, Edgware

Jake Rothman replies:
I thank Godfrey for his extra information, and I have appreciated his letters to Practical Wireless *in the past. The photo of the GET1 point-contact transistor is an especially useful addition, since I couldn't get one. I did try making a point-contact transistor with a couple of needles probing a germanium slice recovered from a shattered OA10 diode. I think I got*

a gain of four for a few minutes before contamination killed it.

Interesting to see the MAT100 in its packet. Maybe we could build a Sinclair TR750 as discussed in John Ellis's letter? I'm surprised it has an ft of 60 MHz, but I couldn't see that it had anything to make it unique for an audio application.

The valve hearing aid is fascinating – so small! I imagined a metal chassis weighing a kilo with an accompanying 90V high-tension battery the size of a large book. I had forgotten there was once 12V valve technology. I used the 12V series of valves (made for car radios of the early 1960s when transistors could only do audio) for Theremins during their 1990s craze. I employed the ECH83 triode/heptode valve where the heptode section made an excellent mixer stage with minimal oscillator coupling, giving a pure sinewave bass I've never bettered with solid-state. The final stage of valve development was the Nuvistor – as seen in Godfrey's photo – not much bigger than a TO5 transistor, but much more resistant to high voltage pulses.

Maybe 'older engineers' like ourselves should pool together our collections of historical electronic bits for permanent exhibition, perhaps at MOSI, the Museum of Science and Industry in Manchester. It has some of the best examples I've seen and they are normally switched on.

I'm glad Godfrey fixed his tumble dryer, so much better to repair than crush and recycle. I often use motor-start capacitors for audio work. Most have a polypropylene dielectric, which is considered one of the lowest distortion materials. I recently bought some 100µF 600V units from Colin Rich at Breconjess, which made very effective tweeter protectors for a PA system I built. These capacitors prevent low-frequency thumps and hum bursts from damaging the delicate two-inch horn drivers. They are also useful for high-tension filtering in valve circuitry, avoiding the use of electrolytic capacitors, which regularly dry out with time and often cost more. Motor-start capacitors do tend to be very big, but this is not a problem when used in huge loudspeaker cabinets or valve equipment.

Getting back into electronics

Dear editor

I have worked in many different electronic disciplines, but have not had 'hands on' experience, including hobby projects, for over 40 years. Somehow, the advent of integrated circuits removed the magic of construction.

Recently, however, having almost reached the age of seventy-nine, I was struck with an urge to build something again. Quite by chance I discovered a copy of your January 2015 issue, on a market stall, containing an article on constructing the *Tiny Tim Amplifier*; this seemed the ideal project to satisfy my construction urge.

How things have changed! I've spent countless hours trying to understand, what is new to me, the various acronyms, trying to source individual components, as firms supplying kits of parts, common in my past, seem to have disappeared – but the biggest stumbling block has proved to be a source for the specified heat-sinks. Neither Farnell nor Radiospares, the two suppliers of Element14 parts, recognise the part number or description quoted in the parts list! Unless you can supply me with your source details I shall have to abandon the project and return to gardening!

Finally, I mentioned that I discovered the magazine on a market stall; I have been unable to find the February issue in my local newsagents' shops (including WHSmith). Also, the editorial speaks of 50 Golden Years of publication, yet I cannot remember seeing or knowing of it before. I do, however, remember *Practical Electronics* and other titles. Is *Everyday Practical Electronics* descended from these?

GJ Gillson, by email

EPE *Online Editor Alan Winstanley replies:*
Most of our projects are sourced from Australia's Silicon Chip *magazine, so component ID queries crop up from time to time. Unfortunately,* EPE *cannot promise 1:1 close technical support, as this would require immense resources on our part. The best place to ask for technical help is in the* EPE Chat Zone *forum –* ***www.chatzones.co.uk*** *– where many readers are very happy to offer support. A section called 'Shop Talk' highlights any late-breaking news or component-sourcing Q&As.*

It is very common these days to source parts by mail order/ Internet from suggested sources in Australia or eBay for example. Sometimes, full kits are available online (eg see Jaycar or Altronics websites). Delivery to the UK is never a problem.

The exact TO-220 12°C/W heatsink appears to be available from Jaycar Australia. Jaycar is a reputable, recommended mail order supplier, who ship constantly to the UK. Go to ***www.jaycar.com.au*** *and search for Cat No HH8504*

Drawings are available from Jaycar to enable users to make a substitution. These are only small TO-220 type power transistor heatsinks and appear to be nothing special. In my experience they can be freestanding, clip-on or some have extra PCB pillars. A wide range is also listed on eBay.

As the Tiny Tim Amplifier *evolved from the earlier* Hi-Fi Stereo Headphone Amplifier *project (October 2014 issue et seq.) I wanted to point out that it is probably much less straightforward to build the* Tiny Tim *'out of the box'. This is not really a cut and dried project, but one for the experienced audio enthusiast who would need to be resourceful and adaptable. The project articles contain some intensive close technical details and unfortunately there is no short-cut to digesting them. It may be useful to refer to the earlier* Headphone Amplifier *project, and also some extra hardware parts are highlighted in* Part 2 *(Feb 2015, page 28). I say this just to explain in advance that some construction head-scratching and puzzling over parts are to be expected!*

EPE *is indeed 50 years old and our November and December 2014 issues contained special supplements that I wrote, charting our heritage from the 1964* Practical Electronics *issue No. 1 to the present day.* Everyday Electronics *was spun out of* PE *to appeal to newcomers. For well-documented reasons both titles went their separate ways before merging to become* EPE*, owned by the present publishers.*

Back issues of EPE *magazine can be bought from us online or over the phone, please see* ***www.epemag.com*** *for details. Your local WHSmith should be able to order a copy for you, but the best (cheapest) way is to order a subscription from us for direct delivery to your door. I'm afraid it's a sign of the times – it is ever harder to get a niche title onto the newsagent shelves. Please contact us if you have difficulties with your local newsagent.*

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All prices include VAT and postage and packing. Add £2 per board for airmail outside of Europe. Remittances should be sent to **The PCB Service, Everyday Practical Electronics, Wimborne Publishing Ltd., 113 Lynwood Drive, Merley, Wimborne, Dorset BH21 1UU. Tel: 01202 880299; Fax 01202 843233; Email: orders@epemag.wimborne.co.uk. On-line Shop: www.epemag.com**. Cheques should be crossed and made payable to *Everyday Practical Electronics* **(Payment in £ sterling only)**.

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Next Month

Content may be subject to change

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Well, you've asked for it, and here it is – a high quality, practical valve amplifier. Tried and tested, the *Currawong Amplifier's* circuit has been designed for components readily available in 2015. Each channel uses two 12AX7 twin triodes for the preamp and phase splitter stages, and two 6L6 beam power tetrodes in the class-AB ultra-linear output stage. It performs very well, with low distortion and noise.

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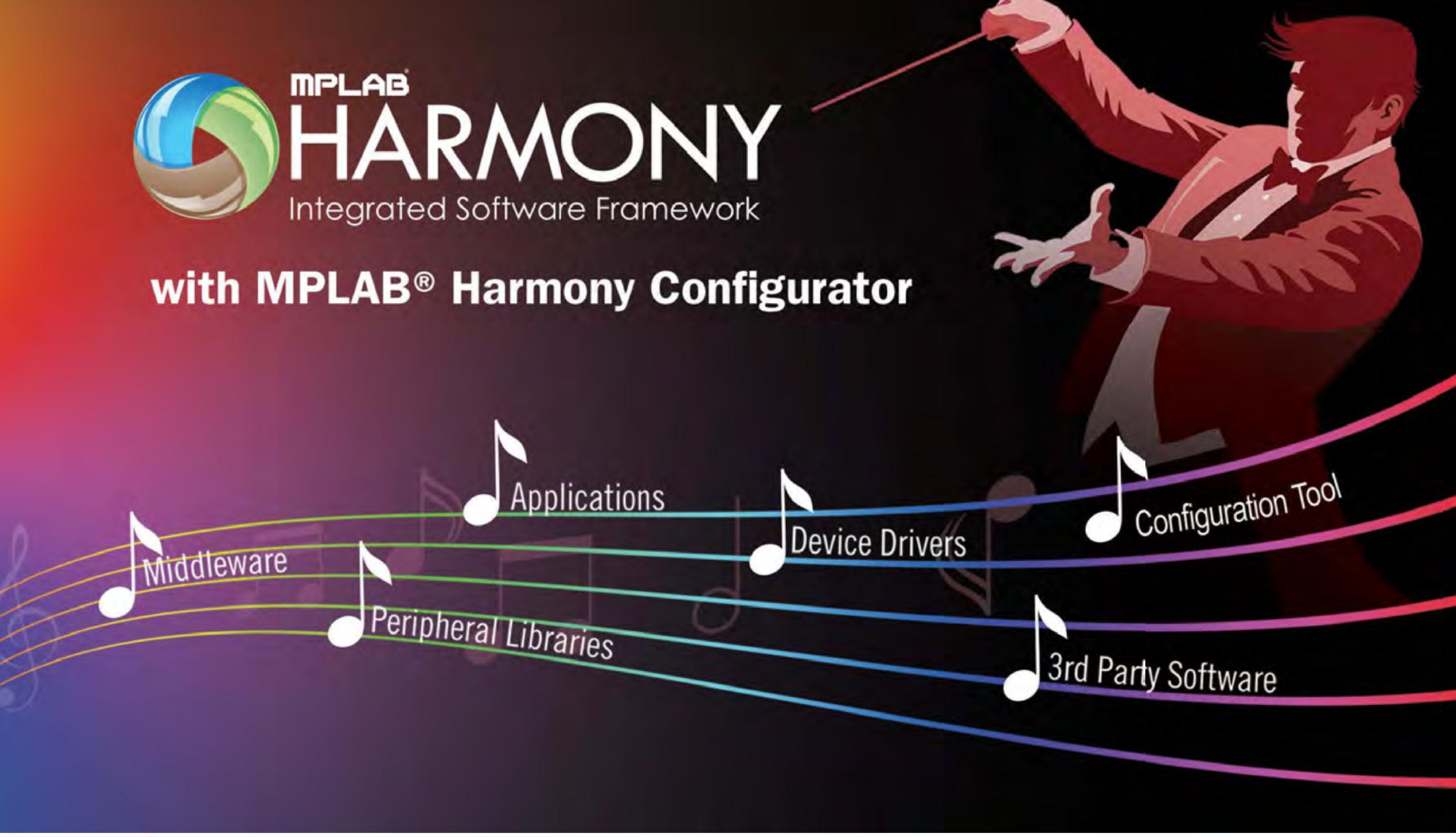
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